



1987 Analog Data Acquisition Applications Seminar

A/D Converters

D/A Converters

Voltage References

Op Amps/Buffers

Power Supply Circuits

Analog Switches

Analog Multiplexers

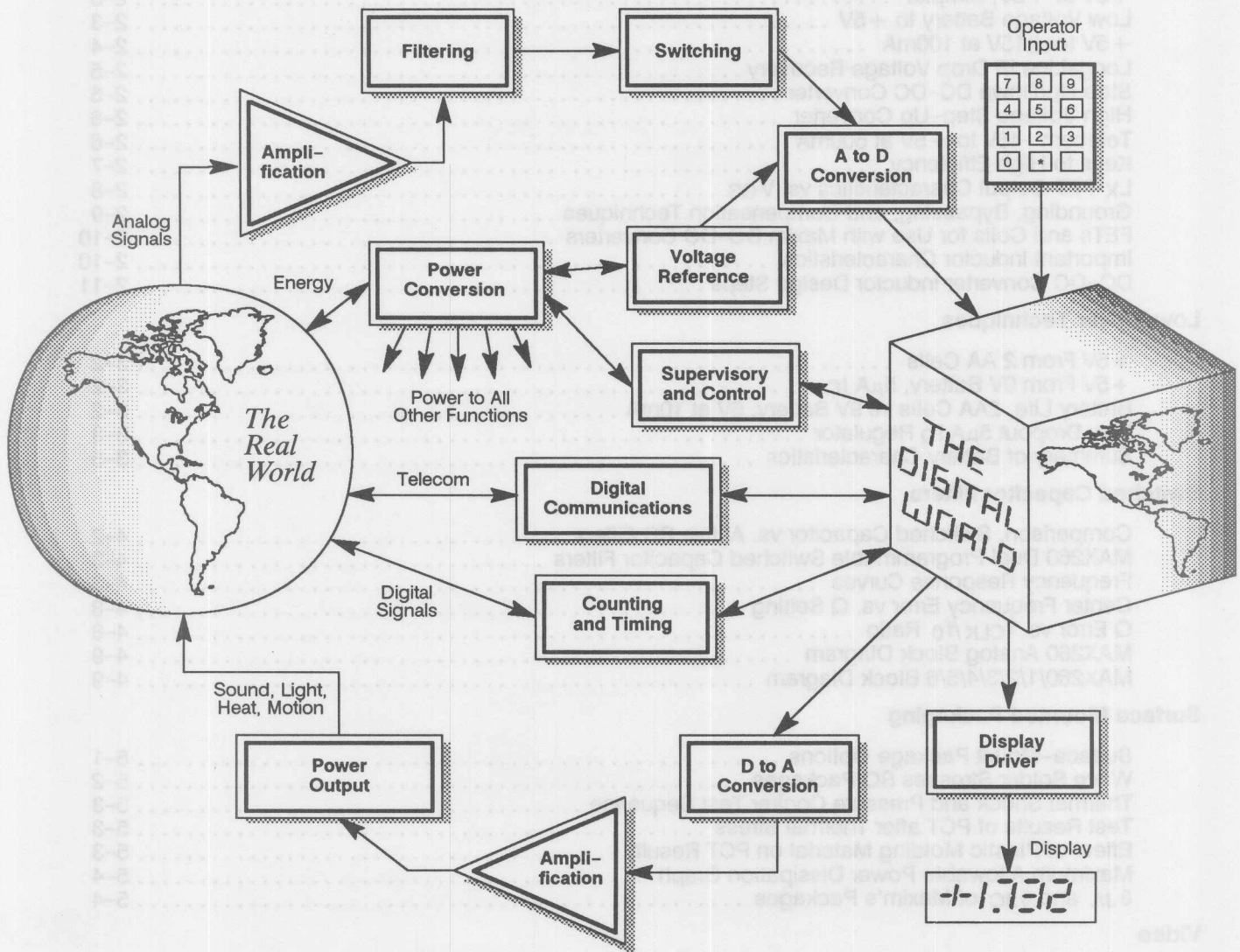
Filters/Interface Circuits

Display Drivers

Timers/Counters

MAXIM

The Connection Between Worlds



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MAXIM'S CHARTER

To Be a Full Line Supplier of Analog and Data Acquisition Products

MAXIM'S PRODUCT LINE

- | | |
|-------------------------|------------------------------|
| • A/D Converters | • Analog Switches |
| • D/A Converters | • Analog Multiplexers |
| • Voltage References | • Filters/Interface Circuits |
| • Op Amps/Buffers | • Display Drivers |
| • Power Supply Circuits | • Timers/Counters |

MAXIM TECHNOLOGIES

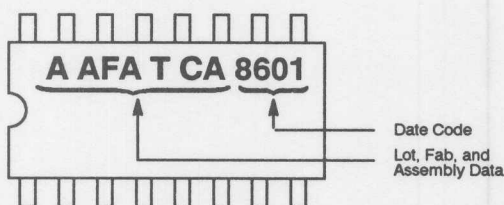
- Silicon Gate CMOS
- 20V Metal Gate CMOS
- 40V Metal Gate CMOS
- 120V Silicon Gate CMOS
- 44V Bipolar
- Hybrid
- Thin Film
- Laser Trimming

QUALITY FROM THE BEGINNING

- Analytically Derived Design Rules
- Improved ESD and SCR Latch-up Protection
- Full Parametric Tests on Incoming Wafers
- All Devices Tested at Maximum Rated Temperature
- 100% Burn-In is Standard
- Full Parametric Testing After Burn-In
- QA Reviews Lot History Before Any Product is Shipped

TRACEABILITY

All Maxim Products Are Fully
 Traceable From Individual Product
 Back To Starting Material



ENVIRONMENTAL TESTING

Pressure Pot

- Every Assembly Lot Sampled (Non-Hermetic Packages)
- Provides Control Gate for Integrity of Passivation and Assembly

85°C/85% Humidity

- Every Generic Device Type Evaluated on Regular Basis
- Test time is 1000 Hrs., with Extensions to 2000 Hrs.
- Excellent Correlation Between Pressure Pot & 85/85 Testing

ACCELERATED LIFE TESTING

- More Than 145,000 Devices Tested
- Every Manufacturing Lot Represented
- 192 Hours at 150°C
- Individual Product Reports Available

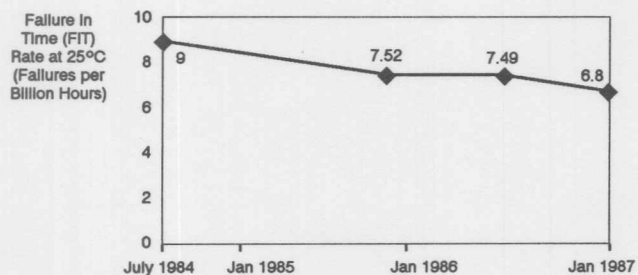
6.8 Failures Per Billion Device Hours

MAXIM'S RELIABILITY RESULTS

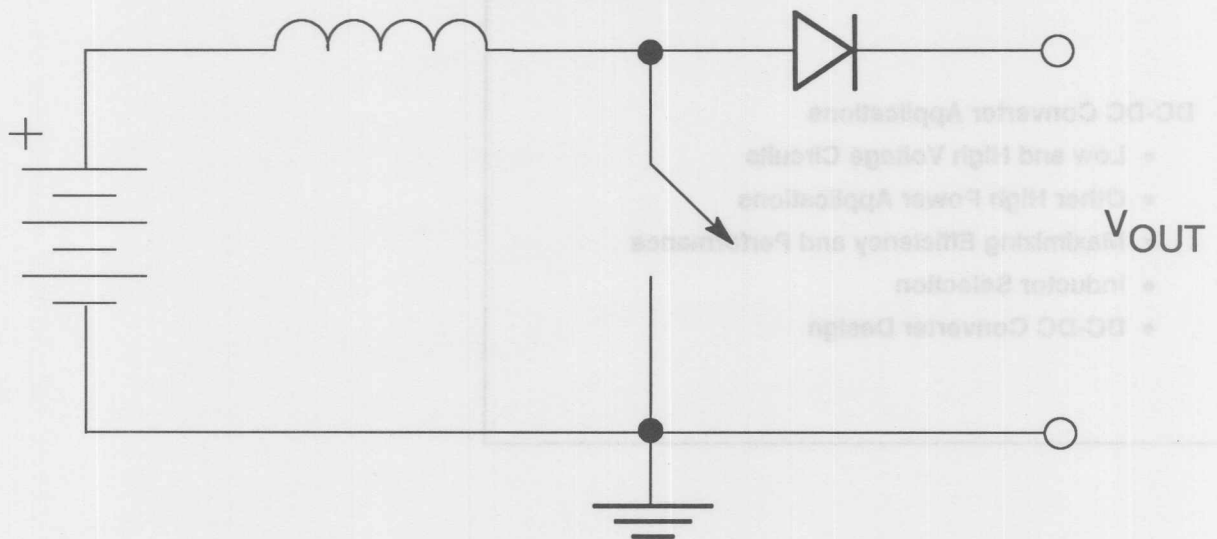
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**FAILURES IN ONE
BILLION HOURS**

MAXIM'S RELIABILITY HISTORY



DC - DC CONVERTERS



DC-DC Converters

DEVICE	DESCRIPTION	INPUT VOLTAGE	OUTPUT VOLTAGE	COMMENTS
ICL7660	Charge Pump Voltage Inverter	1.5V to 10V	$-V_{IN}$	Not regulated
MAX4193	DC-DC Boost Converter	2.4V to 16.5V	$V_{OUT} > V_{IN}$	RC4193 2nd source
MAX630	DC-DC Boost Converter	2.0V to 16.5V	$V_{OUT} > V_{IN}$	Improved RC4191 2nd source
MAX631	DC-DC Boost Converter	1.5V to 5.6V	+5V	Only 2 external components
MAX632	DC-DC Boost Converter	1.5V to 12.6V	+12V	Only 2 external components
MAX633	DC-DC Boost Converter	1.5V to 15.6V	+15V	Only 2 external components
MAX4391	DC-DC Voltage Inverter	4V to 16.5V	up to -20V	RC4391 2nd source
MAX634	DC-DC Voltage Inverter	2V to 16.5V	up to -20V	Improved RC4391 2nd source
MAX635	DC-DC Voltage Inverter	2V to 16.5V	-5V	Only 3 external components
MAX636	DC-DC Voltage Inverter	2V to 16.5V	-12V	Only 3 external components
MAX637	DC-DC Voltage Inverter	2V to 16.5V	-15V	Only 3 external components
MAX638	DC-DC Voltage Stepdown	3V to 16.5V	$V_{OUT} < V_{IN}$	Only 3 external components
MAX641	High Power Boost Converter	1.5V to 5.6V	+5V	Drives external MOSFET 10 Watt output
MAX642	High Power Boost Converter	1.5V to 12.6V	+12V	Drives external MOSFET 10 Watt output
MAX643	High Power Boost Converter	1.5V to 15.6V	+15V	Drives external MOSFET 10 Watt output
MAX680	Charge Pump Voltage Quadrupler	2.0V to 6.0V	$+2V_{IN}, -2V_{IN}$	4 external capacitors

MAXIM
ADVANTAGE

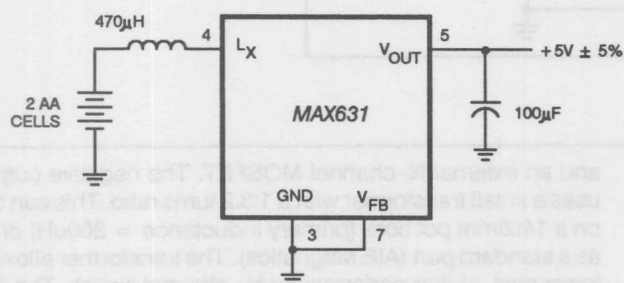
DC-DC Converter Applications

- Low and High Voltage Circuits
- Other High Power Applications
- Maximizing Efficiency and Performance
- Inductor Selection
- DC-DC Converter Design

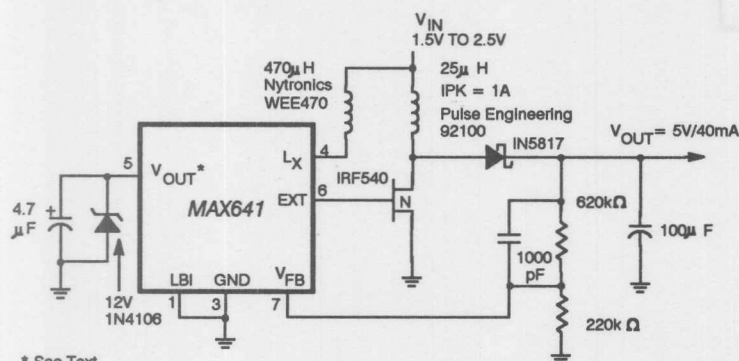
Maxim's CMOS DC-DC Voltage Converters MAX630 through MAX643

- Step-Up, Step-Down, and Inverting
- Minimum External Components
- Small 8-pin Packages
- Preset or Adjustable Output Voltages
- Low Operating Current – 125uA typ.
- On-Chip Low Battery Detection
- Up to 90% Conversion Efficiency

3 VOLTS TO 5 VOLTS



LOW VOLTAGE BATTERY TO +5V



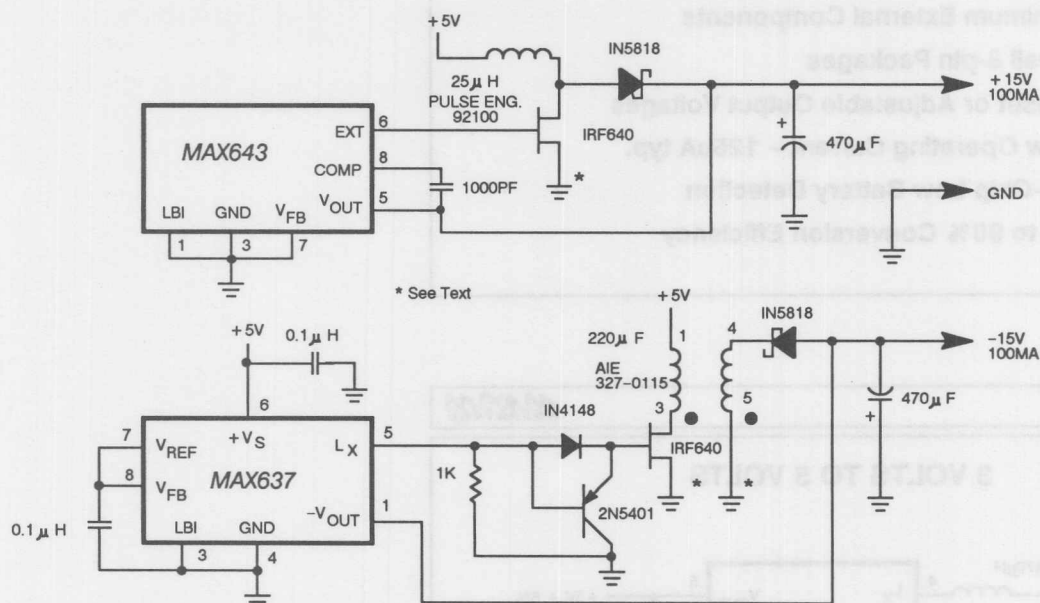
* See Text

By connecting a second small inductor to the L_X output of a MAX641 step-up DC-DC converter, the efficiency and power handling ability of many low voltage converter designs can be dramatically improved. This circuit supplies 5V at 40mA with only a 1.5V input.

The second coil (470uH) works with L_X to form a second stepup converter whose only function is to supply power to the MAX641 chip. An internal diode steers the coil's discharge current to to a filter capacitor and zener clamp at V_{OUT} . This way the MAX641 operates from from 12V and provides a larger gate drive signal to the MOSFET, which in turn has lower on resistance than if 5V were connected to V_{OUT} .

* V_{OUT} is actually the MAX641's voltage input and not an output per se. The pin is labeled this way because it usually connects to the circuit output to provide power and the feedback signal back to the chip when the MAX641 is used in its standard configuration. When a MAX64x or MAX63X series device is used in other than the basic configurations, such as here, the V_{OUT} pin is frequently NOT the output of the DC-DC circuit.

+ 5V TO $\pm 15V$ AT 100mA



With a MAX643 step-up and a MAX637 inverting DC-DC converter, sufficient power can be generated from a +5V supply to operate a significant amount of analog circuitry at $\pm 15V$. Since fixed output devices are used, no feedback resistors are needed even though both outputs are regulated.

The positive output uses the standard configuration for the MAX643

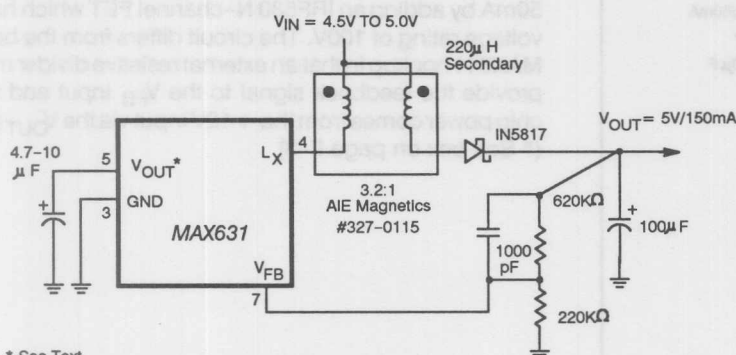
and an external N-channel MOSFET. The negative output circuit uses a small transformer with a 1:3.2 turns ratio. This can be wound on a 14x8mm pot core (primary inductance = 200uH) or obtained as a standard part (AIE Magnetics). The transformer allows use of a lower cost, higher performance N-channel switch. The 2N5401 at the MOSFET gate helps L_X turn the FET off more rapidly for higher efficiency.

*The quality of ground connections is key to the performance of most DC-DC converters. Since the peak current in an inductor or switch ground can reach 1 Amp, these points should have very low impedance paths to supply common. In the best case, separate grounds are used for high current paths than for chip power or feedback connections.

The second coil (L₂) works with L₁ to form a resonant step-up converter which only functions to supply power to the MAX643 chip. An internal diode allows the coil's discharge current to flow to a filter capacitor and return to the MAX643. The MAX643 operates from 15V at 100mA. The MAX643 provides a large signal to the MOSFET, which in turn has a low on-resistance that is 5V when connected to V_{OUT}.

V_{OUT} is actually the MAX643's voltage input and not an output pin. The pin is labeled this way because it was only connected to the output to provide power and the feedback signal back to the chip when the MAX643 is used in its standard configuration. When a MAX643 or MAX644 device is used in other than the standard configuration, such as here, the V_{OUT} pin is actually NOT the output of the DC-DC circuit.

LONG-LINE IR DROP VOLTAGE RECOVERY



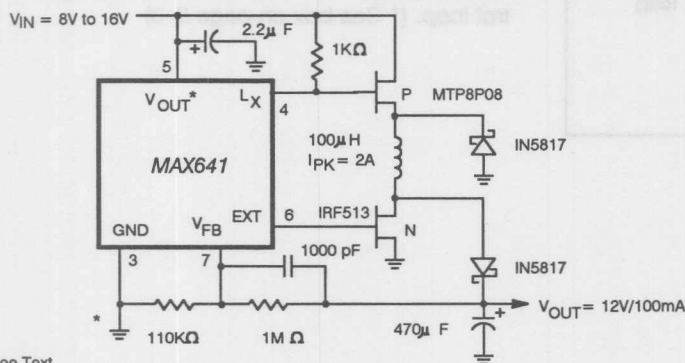
* See Text

This circuit provides a unique solution to a common system-level power distribution problem: When the supply voltage to a remote board must traverse a long cable, the voltage at end of the line sometimes drops to unacceptable levels. This "+5V to +5V" converter, addresses this by taking the reduced voltage at the end of the supply line and boosting it back to +5V. The can be especially useful in remote display devices such as some point-of-sale (POS) terminals where several meters of cable may separate the terminal from the readout.

A MAX631 and a small transformer restore the 5V to 4.5V input back to 5V. The 3.2:1 turns ratio of the transformer allows the MAX631 to provide more than its usual output current, without an external MOSFET, at these relatively low operating voltages. Output current is 5V at 150mA with a 4.5V input.

* The MAX631 also makes use of the reflected voltage in the transformer primary to generate a higher supply voltage of about +9V for itself at V_{OUT} . By operating at 9V rather than 5V, the on resistance of L_X is reduced. (Also see box on page 2-3)

STEP UP/DOWN DC-DC CONVERTER

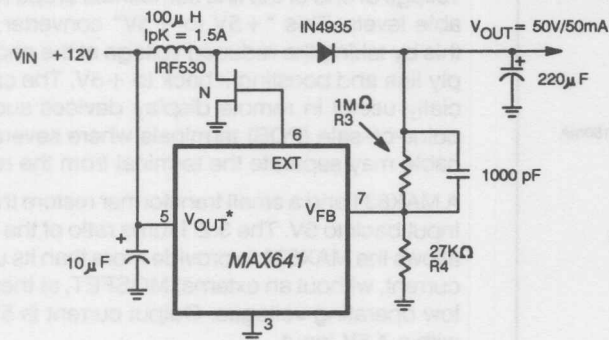


* See Text

Positive output step-up and step-down DC-DC converters have a common limitation in that neither can handle input voltages which may be both greater than or less than the output. For example, when converting a 12V sealed lead acid battery to a regulated +12V output, the battery voltage may vary from a high of 15V down to 10V.

By using a MAX641 to drive separate P- and N-channel MOSFETs, both ends of the inductor are switched to allow noninverting buck/boost operation. A second advantage of the circuit over most boost-only designs is that the output goes to 0V when SHUTDOWN is activated. A drawback is that efficiency is not optimum because 2 MOSFETs and 2 diodes increase the losses in the charge and discharge path of the inductor. The circuit delivers +12V at 100mA at 70% efficiency with an 8V input. (* See boxes on page 2-3 and 2-4)

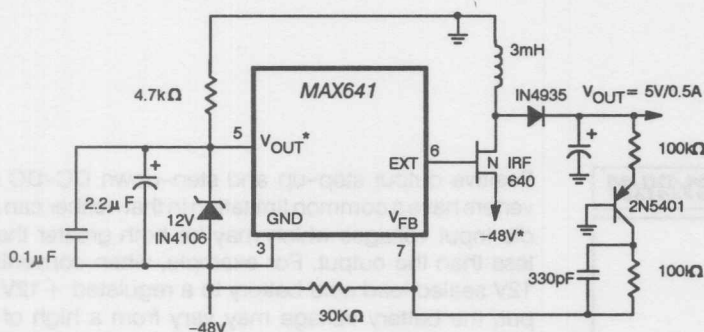
HIGH VOLTAGE STEP-UP CONVERTER



* See Text

The output voltage limits of the MAX6XX series DC-DC converters can be exceeded once an external FET or transistor with an adequate voltage rating is used as the switch. Here a +12V input is converted to +50V at 50mA by adding an IRF530 N-channel FET which has a voltage rating of 100V. The circuit differs from the basic MAX641 hookup in that an external resistive divider must provide the feedback signal to the V_{FB} input and that chip power comes from the +12V input via the V_{OUT} pin. (* See box on page 2-3)

TELECOM -48V TO +5V AT 0.5A



* See Text

The small current consumption of a MAX641 allows it to be biased at a -50V rail with a shunt zener diode so that it can convert -50V to +5V. This is a common requirement in telecom systems where logic circuitry must be powered from the central office battery voltage.

A small high voltage PNP transistor level shifts the a feedback signal from the +5V output down to the MAX641, whose ground pin (pin 3) is tied to the -50V input. The chip is biased this way so that EXT can directly drive an N-channel MOSFET to switch the inductor to -50V. This way the circuit operates much like a step-up DC-DC converter. The 330pF capacitor provides feed-forward compensation to stabilize the regulator's control loop. (* See box on page 2-3)

KEYS TO EFFICIENCY

- Switch Resistance
- Inductor Series Resistance
- Inductor Core Losses
- Diode Forward Voltage
- Quiescent Current
- Switching Time

CMOS DC-DC converters are generally used in low power applications where efficiency is of prime importance. This table lists the major causes of energy loss. Methods of optimizing each are listed below:

Switch Resistance—Use highest possible chip operating voltage (V_{OUT} or $+V_S$). Use external MOSFET or transistor.

Inductor Resistance—Select lowest possible DC resistance.

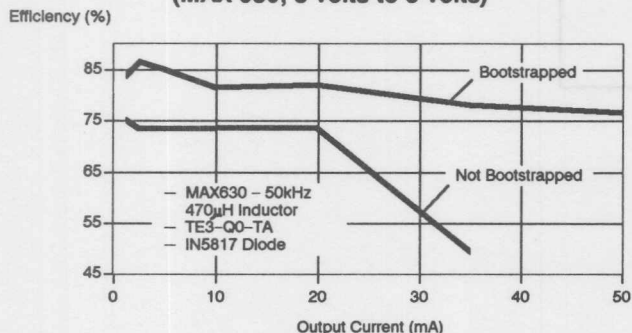
Inductor Core Losses—Select larger inductor size or lower loss core material.

Diode Forward Voltage—Use Schottky diodes for minimum forward voltage.

Quiescent Current—Use low I_Q devices, i.e. MAX6XX DC-DC converters. Select maximum values for feedback and pullup resistors.

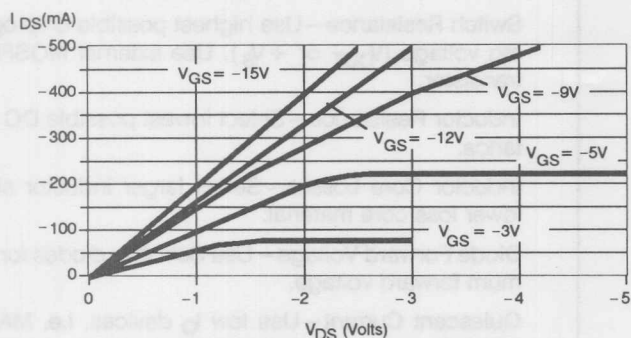
Switching Time—Minimize rise and fall time of MOSFET gate drive. Add external driver for MOSFETs with high gate capacitance. Use high speed steering diodes (1N4001 NOT RECOMMENDED).

EFFICIENCY VS OUTPUT CURRENT (MAX 630, 3 volts to 5 volts)

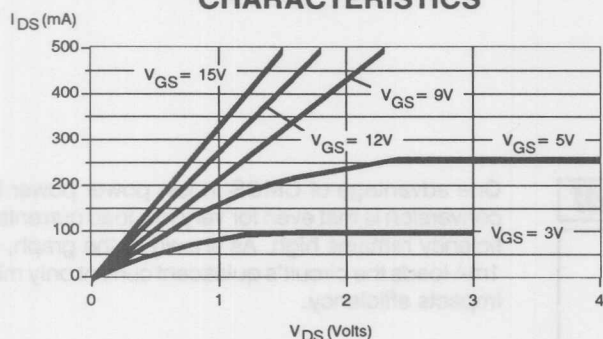


One advantage of CMOS in low power power DC-DC conversion is that even for very low load currents, the efficiency remains high. As shown in the graph, even at 1mA loads the circuit's quiescent current only minimally impacts efficiency.

MAX634-638 L_x OUTPUT CHARACTERISTICS

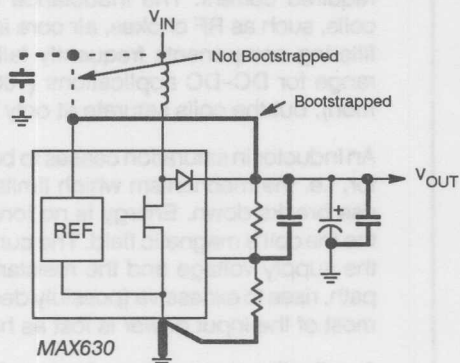


MAX631-633, MAX641-633 L_x OUTPUT CHARACTERISTICS

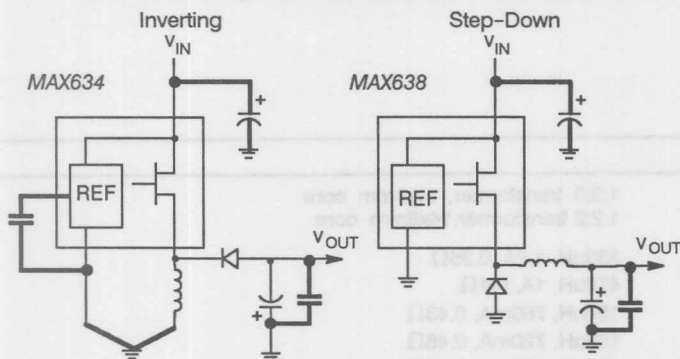


The output characteristics of the internal L_x switches in Maxim's DC-DC converters are shown in these two graphs. The MAX631-633, MAX641-643 plot is for an N-channel output device while the MAX634-638 graph is for a P-channel device. Both graphs show that the saturation current for each device is relatively low and the on resistance is high when the chip's supply voltage (equivalent to V_{GS} in the graphs) is 5V or less. The increase in slope as V_{GS} increases, which corresponds to lower on resistance, shows how conversion efficiency is improved when the chip is powered from the highest available supply voltage.

STEP-UP GROUNDING AND COMPENSATION



DC-DC GROUNDING AND BYPASSING



Loop instability, caused by contaminated ground connections or stray capacitive pickup, can severely limit the performance of an otherwise sound DC-DC converter design. Symptoms of trouble are often high ripple, much poorer than expected efficiency, and "motorboating" or low frequency oscillation. Motorboating occurs

when the control loop of the DC-DC converter outputs pulses in periodic clusters (10-20 pulses long) rather than at random intervals. This happens because the control loop has been destabilized by one or more of the following:

CAUSE

Stray pickup at the feedback node (V_{FB})

Unwanted feedback to the reference

Feedback via GND

CORRECTION

Reduce size of connections at V_{FB} to reduce capacitance to GND.

Add compensation capacitor (typ 1000pF) from V_{FB} or COMP to circuit output.

Bypass reference and/or $+V_S$ to GND (0.1 to 1uF)

Bypass power supply input (V_{OUT} or V_S) with 1 to 10 μ F and 0.1 μ F disc.

Separate high current ground connections from reference, feedback and/or chip GND if possible.

Separate high current supply connections.

These techniques are shown in bold. Not all suggestions apply to all converter types.

IMPORTANT INDUCTOR CHARACTERISTICS

- Maximum Current / Saturation Current
- Series Resistance
- Core Losses
- Physical Size

Not every inductor is appropriate for DC-DC conversion circuitry. Even with the proper inductance value an "unknown" inductor may still saturate if it cannot handle the required current. The inductance of several types of coils, such as RF chokes, air core inductors, and noise filtering components frequently fall in the appropriate range for DC-DC applications (100 to 500uH is common), but the coils saturate at only a few milliamps.

An inductor in saturation ceases to behave like an inductor, i.e. the mechanism which limits the rate of current rise breaks down. Energy is no longer being stored in the the coil's magnetic field. The current, limited only by the supply voltage and the resistance in the charging path, rises to excessive (possibly destructive) levels and most of the input power is lost as heat.

A coil will not saturate as long as its charging current does not exceed its current rating. However, in most DC-DC designs the peak inductor current (I_{pk}) is significantly greater than the average output current, often by as much as 4 to 6 times. This "peak" current flows EVERY time the L_X switch turns on, not just during peak loads. I_{pk} may be as high as several Amps in DC-DC circuits using external MOSFET switches and low inductance values (< 100uH).

INDUCTOR LISTING

MFG	PART	
AIE Magnetics	327-0115	1:3:3 transformer, 14x8mm core
	327-0116	1:2:2 transformer, 14x8mm core
Caddell Burns	6860-19	330uH, 1.2A, 0.35Ω
	6860-21	470uH, 1A, 0.4Ω
	7070-27	150uH, 760mA, 0.43Ω
	7070-28	180uH, 720mA, 0.48Ω
	7070-29	220uH, 670mA, 0.55Ω
Dale	IHA-104	500uH, 1A, 0.5Ω
	TE3-Q0-TA	150uH, 1.2Ω, toroid
	TE3-Q0-TA	470uH, 3.3Ω, toroid
	TE3-Q4-TA	1mH, 0.82Ω, toroid
Gowanda	1B103	100uH, 1A, 0.25Ω
	1B253	250uH, 1A, 0.44Ω
	1B503	500uH, 1A, 0.56Ω
	050AT1003	100uH, 1A, 0.05Ω
Pulse Eng.	92100K	25uH, 2.5A, toroid
	92101K	50uH, 2.5A, toroid

This is a partial listing of coils that have been used successfully with MAX series devices in various DC-DC converter circuits. Their per-

formance for the most part is not unique and other equivalent parts can be expected to function equally well.

Below is a partial list of power MOSFETs which are appropriate for use with MAX6XX DC-DC converters when current above the rating of the internal L_x switch is needed.

The ON resistance of most power MOSFETs is specified with 10V

gate drive. This however is not always available in low voltage DC-DC converter designs. Performance with 5V gate drive is now being specified by some manufacturers. In addition, though not guaranteed, many other devices also have reasonable ON resistance with 5V gate drive. Some of these are also listed.

Power MOSFET Listing

PART#	PKG	R _{ON} (Ω) at (I _{DS} , V _{GS} = xV)	MAX V	\$(100)	MFG
N-CHANNEL MOSFETS					
BUZ71A	TO-220	0.1 (6A, V _{GS} = 10V)	50	1.00	MOT/SI/SM
IRF513	TO-220	0.8 (2A, V _{GS} = 10V) 1.2 (1A, V _{GS} = 5V) *	100	1.06	IR/SI/GE
IRF530	TO-220	0.18 (8A, V _{GS} = 10V) 3(4A, V _{GS} = 5V) *	100	1.95	IR/SI/GE
IRF620	TO-220	0.8 (2.5A, V _{GS} = 10V) 1.3 (2.5A, V _{GS} = 5V) *	200	2.26	IR/SI/GE
IRF640	TO-220	0.18 (10A, V _{GS} = 10V)	200	4.00	IR/SI/GE
IRFD121	4p DIP	0.3 (1.3A, V _{GS} = 10V)	60	2.10	IR/GE
RFP12NO8L	TO-220	0.2 (1.3A, V _{GS} = 5V)	80	1.90	RCA
P-CHANNEL MOSFETS					
IRFD9120	4p DIP	0.6 (1.3A, V _{GS} = -10V)	-100	2.77	IR/GE
IRF9620	TO-220	1.5 (3.5A, V _{GS} = -10V)	-200	3.90	IR
RFP6P08	TO-220	0.6 (V _{GS} = -10V)	-100	4.00	RCA
MTP2P45	TO-220	6 (1A, V _{GS} = -10V)	-450	4.30	MOT
MTP8P08	TO-220	0.4 (4A, V _{GS} = -10V)	-80	1.75	MOT

* specification not guaranteed by manufacturer

Manufacturer code: IR = International Rectifier, SI = Siliconix, MOT = Motorola, GE = General Electric, SM = Siemens, RCA = RCA(GE?)

The following text describes design procedures for the three basic configurations implemented with Maxim's CMOS DC-DC converter products.

DC-DC CONVERTER DESIGN STEPS

- Pick
 - V_{IN} Range
 - V_{OUT} and I_{OUT}
- Determine
 - Inductor Value
 - Peak I_L
 - Need For External FET

THE EASY WAY TO LOW POWER DC-DC CONVERTER DESIGN

Working with low power DC-DC converters is a relatively simple chore in most cases. A reasonable "non mathematic" design approach is to modify a finished design that is already close to your

goals. The following tables provide some common designs to start with.

Step-up DC-DC Converters Using MAX631/632/633

V_{IN} (V)	V_{OUT} (V)	I_{OUT} (mA)	Eff. (%)	INDUCTOR		
				P.N. (Note 1)	μH	Ω
2	5	5	78	6860-21	470	0.4
2	5	10	74	1B253	250	0.44
2	5	15	61	1B103	100	0.25
3	5	25	82	6860-21	470	0.4
3	5	40	75	7070-29	220	0.55
3	12	5	79	6860-10	330	0.35
3	12	10	79	7070-28	180	0.48
5	12	12	88	6860-21	470	0.4
5	12	25	87	6860-19	330	0.35
3	15	5	73	7070-29	220	0.55
3	15	8	71	7070-27	150	0.43
5	15	10	85	6860-21	470	0.4
5	15	15	85	6860-19	330	0.35
8	15	35	90	1B503	500	0.56

Inverting DC-DC Converters Using MAX 635/636/637

V_{IN} (V)	V_{OUT} (V)	I_{OUT} (mA)	Eff. (%)	INDUCTOR		
				P.N. (Note 1)	μH	Ω
+3	-5	5	60	6860-19	330	0.35
+5	-5	25	76	6860-19	330	0.35
+9	-5	40	79	6860-19	330	0.35
+12	-5	45	85	6860-21	470	0.40
+15	-5	50	90	6860-23	680	0.55
+5	-12	12	74	6860-19	330	0.35
+9	-12	30	84	6860-19	330	0.35
+12	-12	40	89	6860-21	470	0.40
+3	-15	2	65	6860-21	470	0.40
+5	-15	8	77	6860-19	330	0.35
+9	-15	25	85	6860-19	330	0.35

Note 1: See "Important Inductor Characteristics" section to reference inductor part number.

Some guideline for modifying the above circuits are:

- 1) For more output power, REDUCE the inductance value. The down side of this is an increase in peak inductor current. Be sure that this current doesn't exceed the L_X ratings. If it does, then an external transistor or FET will be necessary. Note that the data sheet rating for peak L_X current (I_{pk}) is NOT THE SAME AS OUTPUT CURRENT. I_{pk} is often as much as four times greater than output current.
- 2) For more efficiency, increase the inductor value. Doing this reduces the output power, but it also reduces the peak inductor current so switches losses are also less.
- 3) Changing the output voltage to other than the available fixed outputs can be done by adding two feedback resistors (see data sheet for each device). Increasing the voltage will of course mean reduced output current if the inductor remains the same.

CALCULATING INDUCTOR VALUE

If the above approach proves inadequate, or for those who like to calculate, a more rigorous approach to inductor selection is provided below. A general procedure is presented first, followed by specific instructions for each converter type.

The proper inductor value for any DC-DC converter depends on the desired output power, the input voltage (or range of input voltage), and the converter's oscillator frequency and duty cycle (50kHz and 50% for most parts). The oscillator timing is important because it determines how long the coil charges during each clock cycle. This timing, along with the input voltage, determines how much energy is stored in the coil. The energy at any given instant is a function of the coil's current and inductance:

$$E_L = L I_{pk}^2 / 2 \text{ where}$$

$$I_{pk} = V_L t_{ON} / L$$

The total power that can be put in (or taken out of) the coil is the energy (E_L) times the clock frequency.

$$P_L = E_L f_0$$

Most often the clock is a 50% duty cycle square wave so:

$$t_{ON} = 1 / (2 f_0)$$

The coil power as a function of input voltage, frequency (duty cycle = 50%), and inductance is then:

$$P_L = V_L^2 / (8 f_0 L), \text{ or in terms of } t_{ON}:$$

$$P_L = V_L^2 t_{ON} / (4 L)$$

In step-up and inverting converters, the charging voltage for the coil (V_L) is usually the same as the input voltage (V_{IN}) (ignoring losses).

In step-down converters, $V_L = V_{IN} - V_{OUT}$ because the coil is connected between the input and output voltage when charging.

The above equations generally describe the design of MAX6XX DC-DC converter circuits. The following sections contain more specific design steps for each converter type. In each section, the key equations for inductor value, L , and peak inductor current, I_{pk} , are in bold type.

STEP-UP DC-DC CONVERTER DESIGN

(MAX630/31/32/33/41/42/43)

First choose V_{OUT} , I_{OUT} , $V_{IN}(\min)$, and $V_{IN}(\max)$. Remember that in a step-up converter, V_{IN} must be less than V_{OUT} .

$V_{IN}(\min)$ and $V_{IN}(\max)$ cover the input voltage range, such as beginning and end of life battery voltages. The output power, P_{OUT} , is of course $V_{OUT} \times I_{OUT}$, but the converter also has to make up for losses in the inductor, L_X switch, and catch diode. These losses typically add 10 to 25% to the required input power.

In a step-up converter, power is supplied both via the inductor and directly from the input. This is because V_{IN} remains connected to the coil as it supplies current to the output. Therefore:

$$1. \quad P_{OUT} = P_L + V_{IN} I_{OUT}$$

The power, P_L , that we need from the inductor is then:

$$2. \quad P_L = (V_{OUT} - V_{IN} + V_L) I_{OUT}$$

$V_D \times I_{OUT}$ accounts for losses in the catch diode. Schottky diodes minimize (1N5817) this loss which is significant in low voltage circuits. In high voltage circuits ($V_{OUT} > 10V$) or if efficiency is not critical, signal diodes such as 1N4148 perform well if their reverse voltage and current rating is not exceeded.

In order to get P_L out of the inductor, that much power must be put in. In an ideal system, i.e. minimal switch losses, Eq. 3 or 4 provides the power in the coil:

$$3. \quad P_L = V_L^2 / (8 f_0 L), \text{ or in terms of } t_{ON}:$$

$$4. \quad P_L = V_L^2 t_{ON} / (4 L)$$

Solving for the inductor value, L , and substituting Eq. 4 for P_L :

$$5. \quad L = V_{IN}^2 / (8 f_0 I_{OUT} (V_{OUT} + V_D - V_{IN}))$$

f_0 is the DC-DC converter's clock frequency. Eq. 10 assumes that f_0 is a square wave with a 50% duty cycle. On the MAX630 and MAX4193 the clock frequency can be adjusted. It is preset to a fixed rate (50kHz) on the MAX631/32/33 and cannot be changed. In Eq. 10, the MINIMUM expected values should be used for V_{IN} to ensure that there is adequate power under all input conditions.

Besides inductance value, the selected coil must also be rated for the current that it must handle. The peak current that the coil sees is:

$$7. \quad I_{pk} = V_{IN} t_{ON} / L = V_{IN} / (2 f_0 L)$$

t_{ON} is the coil charging time (for one clock cycle) which is equivalent to one half of one f_0 clock period.

When calculating I_{pk} with Eq. 7, the largest expected value of V_{IN} ($V_{IN}(\max)$) should be used so that the maximum current under all operating conditions will be considered. I_{pk} is then compared with the inductor current rating and the current rating of the L_X switch. I_{pk} should of course be less than these values.

If I_{pk} exceeds the peak current rating of the internal L_X switch in the MAX630/4193 (550mA) or MAX631/2/3 (475mA) then an external

MOSFET or transistor with an adequate current rating must be used. The MAX641/2/3 is best suited for these circuits since it is designed to directly drive a power MOSFET.

STEP-DOWN DC-DC CONVERTER DESIGN

(MAX638)

Choose V_{OUT} , I_{OUT} , $V_{IN (min)}$ and $V_{IN (max)}$. In step-down converters, $V_{IN (min)}$ must be greater than V_{OUT} .

$V_{IN (min)}$ and $V_{IN (max)}$ define the converter's input voltage range, such as an unregulated power supply voltage at high and low line. Output power is $V_{OUT} I_{OUT}$, but the converter must also be able to supply power to make up for losses in the inductor, L_X switch, and catch diode. If an 80% conversion efficiency is assumed, P_{OUT} is multiplied by 1.25 in the equations below.

In a step-down converter such as the MAX638, the output power is the sum of the power supplied via the coil and the power supplied directly from the input voltage. The coil is connected between the input and output voltage when it charges, so charging current flows into the load as well. When the coil discharges, current flows from ground, through the diode and coil, into the load. Total output power is:

$$8. \quad P_{OUT} = P_L + V_{OUT} I_{pk}/4$$

P_L is the power supplied by the coil and $V_{OUT} I_{pk}/4$ is the power supplied directly to the load while the coil charges. I_{pk} is the peak charging current of the Inductor. The above equation assumes that the coil charging current rises linearly from 0 to I_{pk} during one half of each oscillator cycle. The AVERAGE coil charging current is then $I_{pk}/4$.

The peak inductor current is a function of the charging voltage ($V_{IN} - V_{OUT}$), charging time (t_{ON}), and coil inductance (L).

$$9. \quad I_{pk} = (V_{IN} - V_{OUT}) t_{ON} / L$$

$$10. \quad I_{pk} = V_{IN} - V_{OUT} / (2 f_0 L)$$

f_0 is the converter's clock frequency. The coil can charge for at most one half of each clock cycle ($1/2f_0$). By substituting Eq. 10 into Eq. 8, we get:

$$11. \quad P_{OUT} = P_L + V_{OUT} (V_{OUT} - V_{IN}) / (8 f_0 L)$$

In order to get P_L out of the inductor we must put at least that much in. The power that is put in is:

$$12. \quad P_L = (V_{IN} - V_{OUT})^2 / (8 f_0 L)$$

By substituting Eq. 12 into Eq. 11, we get:

$$13. \quad P_{OUT} = V_{IN} (V_{IN} - V_{OUT}) / (8 f_0 L)$$

In terms of L , and by multiplying P_{OUT} by 1.25 to account for losses, we get:

$$14. \quad L = V_{IN} (V_{IN} - V_{OUT}) / (10 f_0 P_{OUT})$$

The minimum expected value should be used for V_{IN} ($V_{IN (min)}$) to ensure that there is adequate output power under all conditions. Be-

sides inductance value, the selected coil must also be rated for the current that it will be handling in the circuit. The peak current that the coil will see is express by Eq. 9. When calculating I_{pk} , $V_{IN (max)}$ should be used so that the maximum current under all operating conditions will be considered. I_{pk} is then compared with the current ratings of the inductor and L_X switch.

If I_{pk} exceeds the peak current rating of the internal L_X switch in the MAX638 (550mA) an external MOSFET or transistor, that is rated for the current, can be used with a MAX641/2/3 or MAX631/2/3 type converter. Although these are called "step-up" regulators they can easily be configured for step-down circuits when using an external MOSFET. The MAX638 may also be used with an external switch, but an inverter is also required.

INVERTING DC-DC CONVERTER DESIGN

(MAX634/35/36/37)

Choose V_{OUT} , I_{OUT} , $V_{IN (min)}$ and $V_{IN (max)}$. In an inverting DC-DC converter, V_{IN} may be greater, equal, or less than V_{OUT} .

Output power is $V_{OUT} I_{OUT}$, but the converter has to supply additional power to make up for losses. These typically add 10 to 25% to the required power, depending on external conditions such as component selection and operating voltage.

In an inverting converter (MAX634/35/36/37) all output power is supplied via the coil. One end of the coil remains grounded when it charges and discharges. The total output power is:

$$15. \quad P_{OUT} = P_L - V_D I_{OUT}$$

P_L is the power supplied by the coil and $V_D I_{OUT}$ is the power lost in the steering diode. In order to get P_L out of the inductor, P_L must be put in:

$$16. \quad P_L = V_{IN}^2 / (8 f_0 L)$$

Solving for the inductor value:

$$17. \quad L = V_{IN}^2 / (8 f_0 I_{OUT} (V_{OUT} + V_D))$$

f_0 is the DC-DC converter's clock frequency and assumes that f_0 is a square wave with a 50% duty cycle. On the MAX634, the clock can be adjusted while the MAX635/36/37 have a preset oscillator that cannot be changed. In Eq. 17, $V_{IN (min)}$ should be used to ensure that there is adequate power under all conditions.

Besides inductance value, the selected coil must also be rated for the current that it must handle. The peak current that the coil sees is:

$$18. \quad I_{pk} = V_{IN} t_{ON} / L = V_{IN} / (2 f_0 L)$$

t_{ON} is the coil charging time (for one clock cycle) which is equivalent to one half of one f_0 clock period.

When calculating I_{pk} with Eq. 18, the largest expected value of V_{IN} ($V_{IN (max)}$) should be used so that the maximum current under all operating conditions will be considered. I_{pk} is then compared with the inductor current rating and the current rating of the L_X switch. If it exceeds the peak current rating of the internal L_X switch in the MAX634 (550mA) or MAX635/36/37 (475mA) then an external MOSFET or transistor must be used.

NOTES

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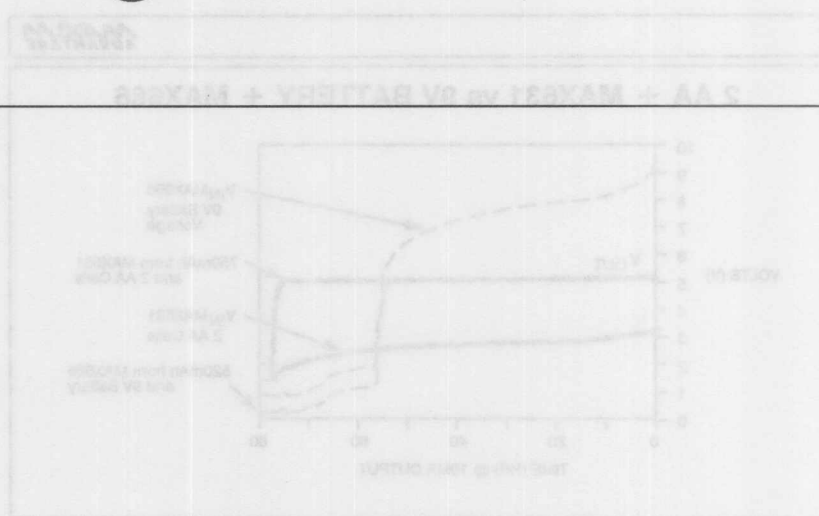
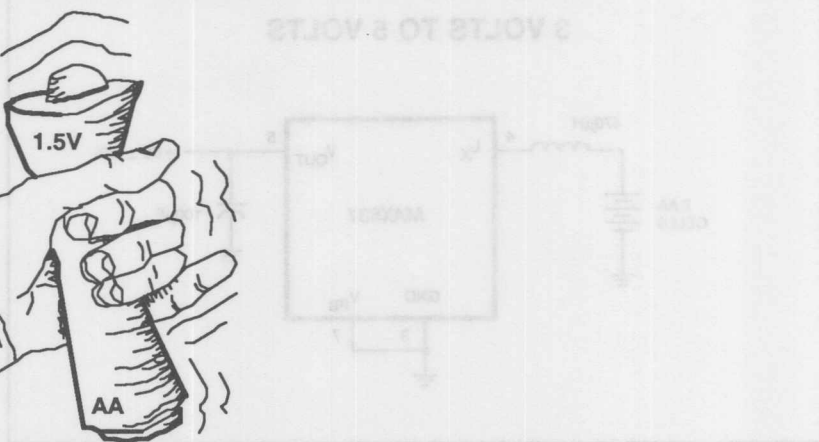
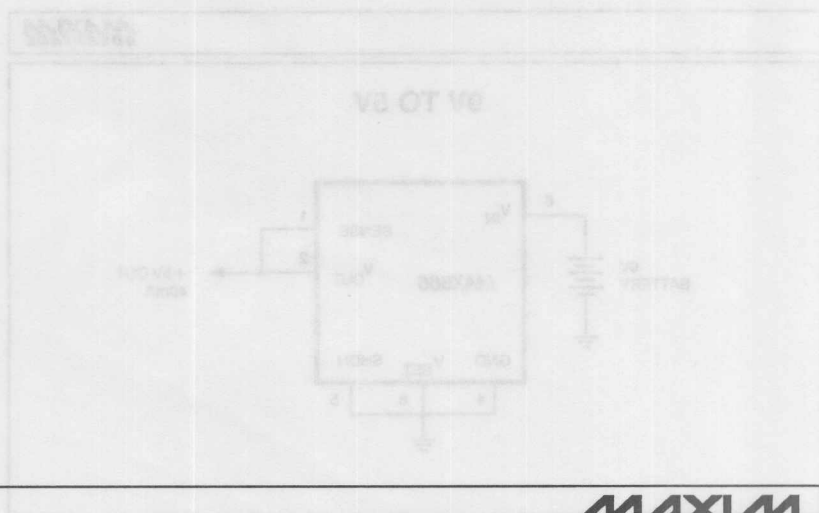
LOW POWER TECHNIQUES



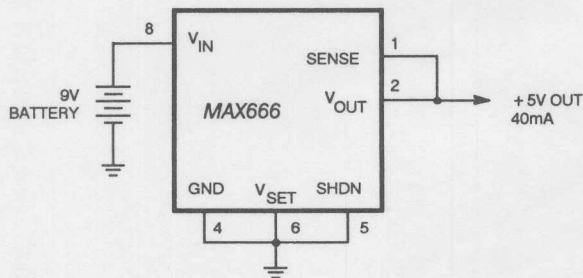
MAX931 DC-DC converter, a small inductor, and two 100kΩ resistors can replace a 9V transistor battery. In a penlight, the MAX931 is used to simultaneously allow a 9V battery to last longer. The AA cells, which contain 20% more energy than the 9V battery, are used to power the MAX931. The MAX931 also has 20% LESS weight and volume. The graph shows a MAX931 and two AA cells delivering 10mA at 5V for a longer time than a MAX931 regulator and a 9V alkaline battery.

MAX931 and a 9V battery then become the primary method of generating 5V, since its quiescent current is only 54μA compared to 100μA with the MAX931. If the operating life of the system is greater than 2000 hours, the MAX931's low I_Q compensates for the lower energy available from the 9V battery.

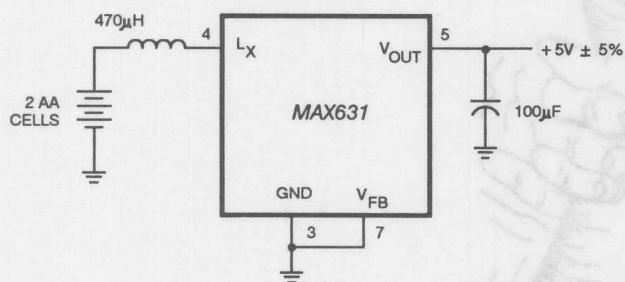
The other information that can be obtained from this graph is the effect of the linear regulator's dropout voltage on battery life. Since the 9V alkaline battery voltage falls rather rapidly near end-of-life, the 0.5V dropout voltage of the MAX931 has only a minimal effect. For this reason, regulators with lower dropout voltage but higher quiescent current than the MAX931 are potentially a poor choice for maximum battery life with alkaline cells.



9V TO 5V

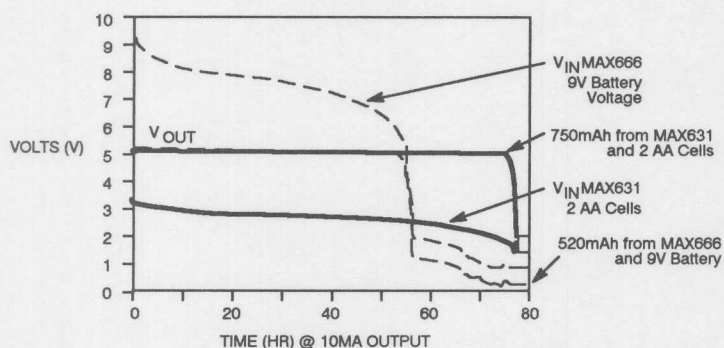


3 VOLTS TO 5 VOLTS



A MAX631 DC-DC converter, a small inductor, and two AA penlight cells can replace a 9V transistor battery, increasing battery life while simultaneously allowing a lower physical profile. The AA cells, which contain 30% more energy than the 6 smaller cells that make up a 9V battery, also have 25% LESS weight and volume. This graph shows a MAX631 and two AA cells delivering 10mA at 5V for a longer time than a MAX666 regulator and a 9V alkaline battery.

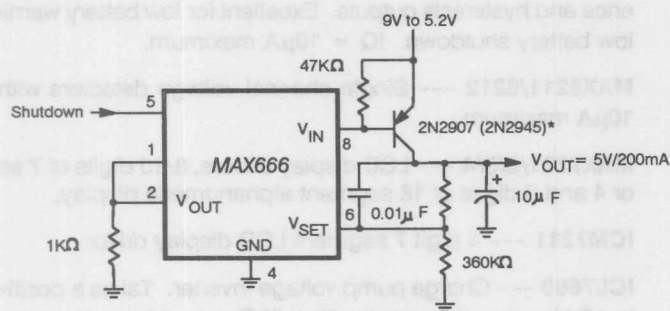
2 AA + MAX631 vs 9V BATTERY + MAX666



If the average current drain is less than 300µA, the MAX666 and a 9V battery then becomes the preferred method of generating 5V, since its quiescent current is only 5µA, compared to 100µA with the MAX631. If the operating life of the system is greater than 2000 hours, the MAX666's low I_Q compensates for the lower energy available from the 9V battery.

The other information that can be obtained from this graph is the effect of the linear regulator's dropout voltage on battery life. Since the 9V alkaline battery voltage falls rather rapidly near end-of-life, the 0.6V dropout voltage of the MAX666 has only a minimal effect. For this reason regulators with lower dropout voltage but higher quiescent current than the MAX666 are frequently a poor tradeoff for maximum battery life with alkaline cells.

LOW DROPOUT LOW I_Q REGULATOR

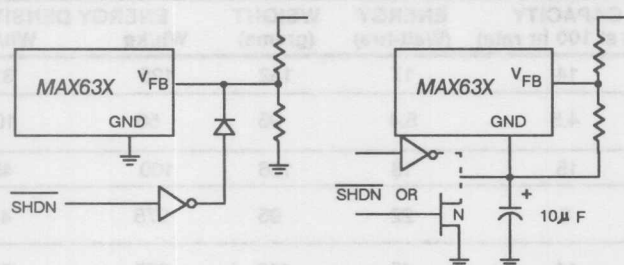


* See Text

A MAX666 or MAX663 CMOS voltage regulator combined with a PNP transistor can supply +5V at several hundred milliamps with as little as 5.3V for an input. Such low dropout performance is beneficial in battery powered products because it extends the useful life of the battery. This is especially true for batteries with sloping discharge curves such as sealed lead-acid and lithium. The quiescent current, which is a function of transistor Beta and load current, is typically 2mA at full load and 10μA at no load. When the circuit is shut down, the current drain is typically 6μA.

The dropout differential can be further reduced by using a 2N2945 as the output transistor rather than a 2N2907. The circuit can then regulate with an input of 5.1V, but the output current is limited to 100mA. At lighter loads, the dropout voltage is even less.

LOW POWER SHUTDOWN TRICKS



Several of Maxim's DC-DC and regulator products include a SHUTDOWN pin which will put the device into a "sleep" mode where as little as 10μA or less is used (varies with device). Due to pin-count and package size limitations, not all products are provided with this feature. At left are some alternate methods of implementing shutdown.

In the left hand circuit, the CMOS inverter overdrives the feedback input (VFB) and fools the DC-DC converter or regulator into thinking its output is too high. The output of the device turns off but the internal circuits remain active. For this reason the shutdown current is higher with this technique than with a true shutdown input.

The right hand technique guarantees near zero (only leakage current flows) quiescent current by lifting the DC-DC converter's ground connection with a CMOS logic gate or a small N-channel FET. It is not appropriate for all applications because a current path may still be left active between the unregulated input and the circuit output in some configurations.

LOW POWER FAVORITES

ICL7611 Family of Op Amps --- Singles, Duals, and Quads with ISupply of only 10 μ A per amplifier. ± 1 V to ± 8 V operation (2V to 16V single supply).

ICL7663 and ICL7664 --- 5 μ A quiescent current voltage regulators for up to 40mA output current. 1.3V to 15V programmable output.

MAX663 --- 5 μ A IQ voltage regulator with preset +5V or externally adjustable output voltage.

MAX666 --- 5 μ A IQ regulator with onboard low battery detector. Fixed +5V or 1.3V to 15V adjustable output.

MAX630-643 --- a complete family of regulating DC-DC converters optimized for battery powered equipment.

ICL7665 --- Dual over/undervoltage detector with onboard reference and hysteresis outputs. Excellent for low battery warning and low battery shutdown. IQ = 10 μ A maximum.

MAX8211/8212 --- Single channel voltage detectors with IQ = 10 μ A maximum.

MAX7231/2/3/4 --- LCD display drivers, 8/10 digits of 7 segment or 4 and 5 digits of 18 segment alphanumeric display.

ICM7211 --- 4 digit 7 segment LCD display driver

ICL7660 --- Charge pump voltage inverter. Takes a positive 1.5V to 10V input and inverts it with a 55 Ω output impedance.

MAX680 --- Charge pump voltage doubler and voltage inverter. Converts +3V to ± 6 V, +5V to ± 10 V.

ICL7136 --- 3-1/2 digit A/D converter with LCD driver uses only 100 μ A maximum supply current from a 9V battery.

Table 1. Performance of Primary (Non-Rechargeable) Battery Chemistries

CHEMISTRY ANODE/CATHODE/ELECTROLYTE	OPEN CIRCUIT V	LOADED V	CAPACITY (Ah at 100 hr rate)	ENERGY (Watt-hrs)	WEIGHT (grams)	ENERGY DENSITY Wh/kg	ENERGY DENSITY Wh/Liter
Alkaline Zn/MnO ₂ /KOH	1.58	1.5 - 1.1	14	17	132	125	315
Carbon Zinc (LeClanche) Zn/MnO ₂ /NH ₄ Cl, ZnCl ₂	1.55	1.5 - 1.0	4.5	5.4	95	55	100
Mercury Zn/HgO/KOH	1.35	1.3	15	18	166	100	450
Lithium Sulfur Dioxide Li/SO ₂ /LiBr	3.0	2.8 - 2.7	8	22	95	275	440
Lithium Thionyl Chloride Li/SOCl ₂ /LiAlCl ₄	3.6	3.5 - 3/4	14	45	113	375	850
*Lithium poly-carbonmonofluoride nLi/(CF) n/ α -butyrolactone	3.0	2.6	10	26	94	275	500
*Lithium Manganese Dioxide Li/MnO ₂	3.25	3.0 - 2.0	14	26	130	230	500

* Extrapolated from data for smaller cells.

Table 2. Secondary or Rechargeable Chemistries

CHEMISTRY ANODE/CATHODE/ELECTROLYTE	OPEN CIRCUIT V	LOADED V	CAPACITY (Ah at 100 hr rate)	ENERGY (Watt-hrs)	WEIGHT (grams)	ENERGY DENSITY Wh/kg	ENERGY DENSITY Wh/Liter
Nickel Cadmium Ni/Cd/KOH	1.35	1.25	4.4	5.3	140	40	120
Lead Acid PbO ₂ /Pb/H ₂ SO ₄	2.1	2.0	2.7	5.4	182	30	100

Primary Battery Chemistries

Dry Cells. There are three popular cell types based on zinc and carbon: the standard LeClanche dry cell, the "heavy duty" zinc chloride cell, and the alkaline manganese cell. As shown in Table 1, there is about a 3 to 1 difference in capacity between the alkaline and LeClanche at the 100 hour discharge rate. The zinc chloride battery falls about midway between the other two. The most significant difference between the characteristics of these three chemistries is their response to high load currents, and their response to continuous versus pulsed applications. In very low drain applications such as 1mA from a D cell, the LeClanche cell has about 1/2 the capacity of the alkaline (7Ah vs 16Ah). When the load current is increased to 80mA, the duty cycle begins to play an important role. The alkaline battery loses virtually no capacity when operated continuously at 80mA, but the LeClanche cell capacity drops from 4Ah to 2.3Ah as the duty cycle of the 80mA load is increased from 1h/day to 24h/day. The zinc chloride cell has 6.2Ah and 6.0Ah capacity under 80mA load, 1h/day and 24h/day respectively.

Summary: The alkaline cell has about twice the capacity of a LeClanche under very light loads, and is over 6 times better under continuous duty high current loads.

Mercury. This cell has a very stable output voltage of about 1.35V, except for those cells which have a small amount of manganese dioxide added, which raises the open circuit voltage to 1.4V. It has good storage characteristics, retaining 85% to 90% capacity after two years at 20°C, and about 80% after one year of storage at 45°C. Mercury cells are generally not recommended for use below 0°C. The "9V" size 6 cell mercury battery has an output voltage of 7.5 to 8.0V for virtually all of its life, and has about 575mAh capacity. The capacity of mercury batteries is nearly independent of duty cycle, except at very high discharge rates. It is generally more expensive than alkaline cells, which have nearly the same performance, except that the output voltage of alkaline cells changes more as the cell is discharged.

Silver Oxide. Primarily used for miniature button and coin cells such as those used in watches and hearing aids, the silver oxide cell has a flat 1.55V discharge. They are best for light loads of C/50 or less. It retains about 70% of its capacity when operated at 0°C, and about 30% at -20°C.

Lithium Manganese Dioxide. (LiMn_2) There are two basic types of LiMnO_2 cells. A pressed powder cathode is used in low discharge rate cells such as the "coin" cells. Flat and cylindrical cells with higher discharge current ratings use a thin pasted electrode on a supporting grid structure. The open circuit voltage is slightly higher than 3.0V. The output voltage is relatively flat at 2.8V to 3.0V up to the last 20% of the discharge period. Like most lithium chemistries, Li/

MnO_2 cells have very good storage characteristics, with about 85% of capacity remaining after 6 years storage at 20°C.

This cell chemistry is available in sizes from 30mAh to over 1Ah. LiMnO_2 cells do not exhibit the voltage delay phenomena sometimes seen in LiSO_2 and LiSOCl_2 cells.

Lithium Sulfur Dioxide (LiSO_2). This is the most mature of the high energy density, high discharge rate lithium chemistries and has been extensively used in military applications such as sonobouys, radios, and beacons. Early LiSO_2 had some safety problems, but manufacturers now extensively test their designs. Indeed, many data sheets have more information about crush tests, nail-through-the-battery tests, and incineration tests than information about discharge characteristics.

Large LiSO_2 cells with high discharge rate capability have safety devices such as high temperature fusible links, over-pressure safety vents, and time delay fuses to prevent catastrophic rupturing (battery manufacturers apparently do not like the term explosion).

The capacity of LiSO_2 cells is about 25% less than its major competitor, lithium thionyl-chloride (LiSOCl_2). LiSO_2 cells have excellent storage characteristics, and maintain a high percentage of capacity over a wide temperature range. Typical cells sizes range from 300mAh upwards to many tens of ampere hours. The open circuit voltage is about 3.0V, and the voltage during discharge is relatively flat.

Lithium thionyl-chloride (LiSOCl_2). Presently the highest energy density system available, LiSOCl_2 is the choice over LiSO_2 in many new applications. In general, LiSOCl_2 cells have about 1/3 more capacity for a given volume or weight than do LiSO_2 cells, but energy densities vary slightly between manufacturers.

Typical cell capacities range from several hundred mAh through 20Ah, but there are cells up to 8,000 ampere hours available. See "Lithium cells suit high-energy military needs", Don Powers, EDN April 85. The open circuit voltage is 3.9V, and the discharge voltage is a constant 3.5 to 3.9V, depending on the discharge rate. High discharge rate D size cells can be operated up to 10A.

Lithium poly-carbonmonofluoride (LiCFn). These are available in both coin cells and cylindrical cells up to 1.2Ah. LiCFn cells are well suited for CMOS RAM backup applications since the self discharge rate is typically only 0.5% per year of storage at room temperature. The cylindrical cells have a high pulse current capability of 1A from a 1.2Ah 2/3A cell (17mm diameter by 33.5mm high, 13.5 grams).

Secondary Battery Chemistries

The two most popular rechargeable chemistries are lead acid and nickel cadmium.

Nickel Cadmium Batteries

Nickel cadmium cells and batteries are available in sizes from about 10mAh to over 10Ah, with a wide variety of packaging styles. 10mAh to 150mAh batteries designed for memory backup are available with pins for direct mounting to printed circuit board, with up to 5 seconds exposure to a solder wave being allowed. Batteries are also available in consumer outlets for the standard AA, C, D, and 9V sizes. The consumer C and D batteries are really repackaged "sub-C" or Cs cells, and both have the same 1.2Ah rating as industrial grade sub-C cells. Industrial grade C cells have around 2.5Ah capacity, and D cells about 4Ah to 4.8Ah. "9V" Ni-Cd batteries come in both 6 cell and 7 cell versions, with nominal terminal voltages of 7.2V and 8.4V respectively. Since the output discharge characteristics is very flat, even a 7.2V output voltage is in most applications an adequate replacement for the 9.5 to 6V output of a 6 cell LeClanche or alkaline battery. Ni-Cd 9V batteries, though, have only 65mAh to 100mAh capacity, much lower than the 500+ mAh ratings of alkaline 9V batteries.

Sealed C and D cells are typically recharged at the 0.1C rate for 14 hours. Ni-Cd batteries in memory backup applications are usually continuously current trickle-charged at 0.002C to 0.10C (where C is the cell's rated capacity). Standard cells are normally charged at 0.1C for 14 hours, since this allows a very simple charging circuit, batteries can be overcharged at 0.1C for extended periods without drastic reductions in life. Special, fast rate rate batteries can be recharged in 20 minutes with a charge rate of 4C, but high rate chargers must have sophisticated methods of detecting end-of-charge, such as sensing the increase in temperature that occurs when full charge has been reached. Other methods of terminating fast-charge include temperature compensated voltage detection, and detection of the voltage reduction that occurs after full charge has occurred.

Cell capacity is typically reduced to 50% of its initial value after 500 to 1500 deep discharge cycles, or after 5 or more years of continuous trickle charge at 0.005C. NiCd batteries have a self-discharge rate of about 0.5% per day at 23°C when near full charge, and retain 20% to 40% capacity after 5 months. At 30°C, 50% to 80% charge is retained after 30 days.

The open circuit voltage of a fully charged Ni-Cd cell is about 1.3V, and the discharge voltage is $1.24V \pm 100mV$ for virtually the entire discharge cycle. After the cell voltage has fallen to around 1.1V, the rate a change of the cell voltage increases rapidly. Cells may be completely discharged without damage, but reverse charging at more than -200mV for extended periods will permanently damage cells. This reverse charge condition most often occurs in multi-cell batteries in which the capacity of the various cells is not well matched.

Ni-Cd batteries come in both sealed and vented types. A sealed cell is a closed environment, and allows the escape of gas only under abnormal conditions which cause safety vents to open. The vented cell allows gases to escape from the cell during normal operation. Sealed cells are most common in the sizes from D cell and below, while vented cells are used in very large batteries in such applications as engine starting and mobile X-ray equipment. Vented cells must be periodically inspected and the electrolyte replenished.

Lead Acid Batteries

Lead acid cells come in many forms, the most common one being the automobile battery, which is typically a vented lead-acid battery. Smaller cells, such as D and X come in fully sealed packages. Typical sealed lead-acid battery capacity is 2.5Ah (10hr rate) for D cells, and 5Ah for X cells. The open circuit voltage of a lead-acid cell can be used to estimate the charge state. For a Cyclon brand cell from Gates, the 25°C open circuit voltage battery is about 2.18V when fully charged, declining linearly to about 1.98V when 10% of capacity is left.

The minimum voltage allowed during discharge is normally 1.6V, and lead-acid batteries should not be allowed to self-discharge below 1.8V. If allowed to self discharged below 1.8V while in storage, the battery will take longer than normal to recharge and the next discharge cycle cannot deliver the rated capacity. Subsequent cycles, however, will result in an increase in capacity to the rated capacity. As with all batteries, the rate of self discharge is a strong function of temperature. After 5 months of storage at 20°C, the typical battery will have 60% capacity remaining. At 40°C, however, the battery would be fully discharged after 5 months.

Lead-acid batteries have extremely low internal impedance. D cells can deliver up to 100A for short periods of time.

Since the terminal voltage of a lead-acid battery rises sharply as it nears 100% charge, the most common charge circuit is a simple constant voltage supply. A constant 2.35V charger will recharge a battery to 90% within 2 hours, and the battery can be left float charging at 2.35V indefinitely to maintain full charge. If the battery temperature will vary significantly, the float voltage should have a temperature coefficient of -2.5mV/°C/cell. Typical float life is 5 to 8 years at 25°C and 2.35V float voltage, decreasing to 2 years for 2.35V float voltage at 50°C.

For deep discharge service that need a fast cycle time, the charge voltage can be increased to 2.45V to 2.7V for even faster charging, but continuous float charging at more than 2.4V is not recommended. Typical cycle life is 2500 for full discharge, and over 1000 for a 25% depth-of-discharge. It is important to fully recharge lead-acid cells, with a 2.35V optimum for batteries that are cycles once per week, and 2.45V for batteries that are cycled once per day.

NOTES

NOTES

FREQUENCY RANGE - Conventional analog filters are still capable of higher frequency operation than switched capacitor types, however newer devices can handle frequencies to 40MHz.

LOW FREQUENCY PERFORMANCE - Switched capacitor filters have significant advantages when dealing with very low frequencies (to 1Hz) since they do not need external components.

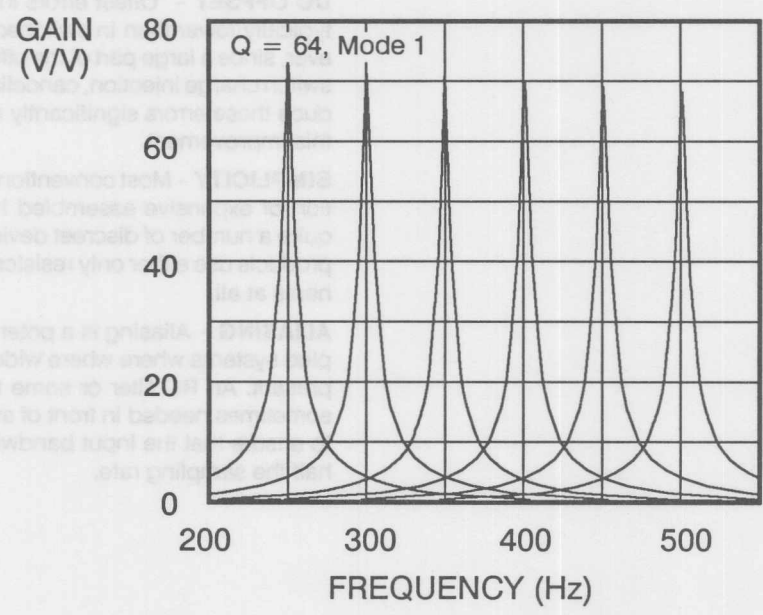
ACCURACY - Switched capacitor filters achieve better than 1% accuracy with ease. Conventional active or passive filters match this performance only with precision external components.

TUNING - Conventional filters are difficult to tune because external components must be varied manually to maintain the filter shape over even a narrow range of frequencies. Switched capacitor filters are easily tuned over a wide range by digital programming of an external clock.

ACTIVE FILTER COMPARISON	
	Switched Cap
Frequency Range	•
Low Freq. Performance	•
Accuracy	•
Tuning	•
Noise	•
DC Offset	•
	Conventional
Frequency Range	•
Low Freq. Performance	•
Accuracy	•
Tuning	•
Noise	•
DC Offset	•



SWITCHED CAPACITOR ACTIVE FILTERS



ACTIVE FILTER COMPARISON

	Switched Cap	Conventional
Frequency Range		●
Low Freq. Performance	●	
Accuracy	●	
Tunability	●	
Noise		●
DC Offset		●
Simplicity	●	

FREQUENCY RANGE – Conventional analog filters are still capable of higher frequency operation than switched capacitor types, however newer devices can handle frequencies to 40kHz.

LOW FREQUENCY PERFORMANCE – Switched capacitor filters have significant advantages when dealing with very low frequencies (to 0.1Hz) since large value external components are not needed.

ACCURACY – Switched cap filters achieve better than 1% accuracy with ease. Conventional active or passive filters match this performance only with precision external components.

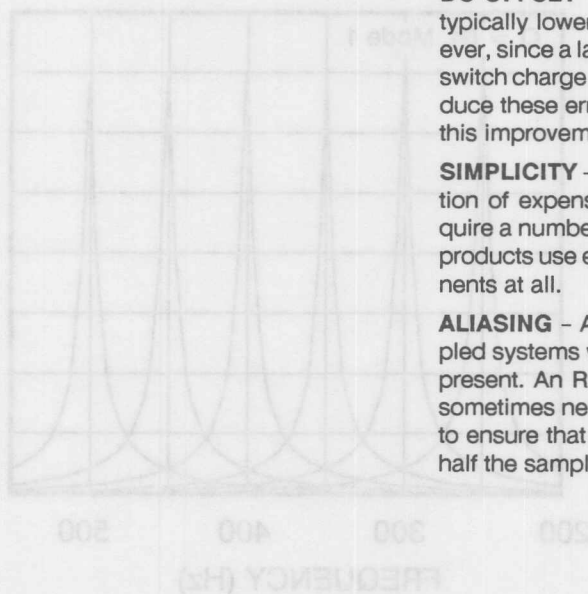
TUNING – Conventional filters are difficult to tune because several external components must be varied simultaneously to maintain the filter shape over even a narrow range of frequencies. Switched cap filters are easily tuned over a wide range by either digital programming or an external clock.

NOISE – Switched cap filters typically exhibit greater noise than conventional active filters. Some clock feedthrough can also be expected. In critical applications a post filter is sometimes used for smoothing.

DC OFFSET – Offset errors in conventional filters are typically lower than in switched capacitor types. However, since a large part of the offset is caused by internal switch charge injection, cancellation techniques can reduce these errors significantly and new designs reflect this improvement.

SIMPLICITY – Most conventional filters, with the exception of expensive assembled hybrids or modules, require a number of discrete devices. Switched capacitor products use either only resistors or no external components at all.

ALIASING – Aliasing is a potential problem in all sampled systems where wide band input signals are present. An RC filter or some type of band limiting is sometimes needed in front of switched capacitor filters to ensure that the input bandwidth remains below one half the sampling rate.



MAX260 FILTER FAMILY

- Dual Second Order Filter
- Bandpass, Lowpass, Highpass, Notch
- Programmed Q and f_O
- μP , Hard Wired, or Resistor Programming
- Crystal, RC Network, or External Clock
- 0.1Hz to 20kHz Center Frequency (f_C) Range
- Military Temperature Grade

SOME ADDITIONAL FEATURES

- 128 programmed Q values
- 64 Programmed f_C 's from CLK/100 to CLK/200
- $\pm 5V$ or Single +5V operation

DUAL PROGRAMMABLE SECOND ORDER ACTIVE FILTERS

MAX260/261	μP Programmed: 6 bit f_O , 7 bit Q, Multi-Input Clock
MAX262/263	Pin Programmed: 5 bit f_O , 7 bit Q
MAX264	Resistor and Pin Programmed, 2 Op-amps, Multi-Input Clock
MAX266	Resistor and Pin Programmed, 1 Op-amp
MAX268/269	Pin Programmed: 5 bit f_O , 7 bit Q, Bandpass Only

The MAX260 family includes eight active filters that allow simplified, minimum component filter designs for a variety of applications. Most products are available in low power versions (odd numbered parts) which maintain bandwidth while reducing supply current from 14 to 2mA. Only output drive capability is reduced in the low power devices.

With the MAX260/261, two independent 2nd order state variable filter blocks can be directly programmed from a μP without using external components.

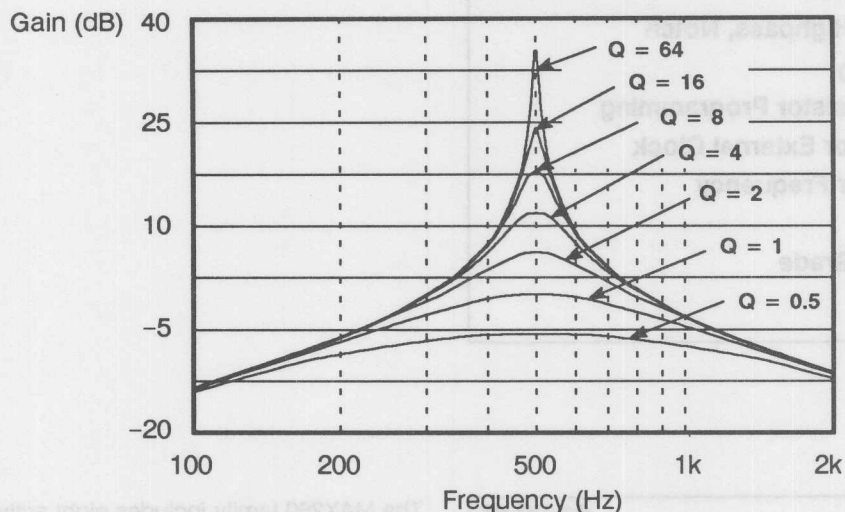
The MAX262/263 is pin programmed and can stand alone without μP support. The MAX264 and MAX266 are programmed via pins strapping and with external resistors for highest trim resolution, which is especially useful in notch filters.

For bandpass only applications, the MAX268/269 provides two 2nd order sections with pin programmed f_C and Q in a 24 pin narrow DIP.

MAX260 Family - MF10 Comparison

MAX260 Family	MF10
No External Components (MAX260-263)	Requires 6 or 8 Resistors
Microprocessor Programmed (MAX260/261) f_O and Q	Resistor Programmed Only
Crystal, RC or External Clock (MAX260/261/264)	External Clock Only
Low Power Versions use 2 mA (MAX261/263/269)	10mA Max. Supply Current

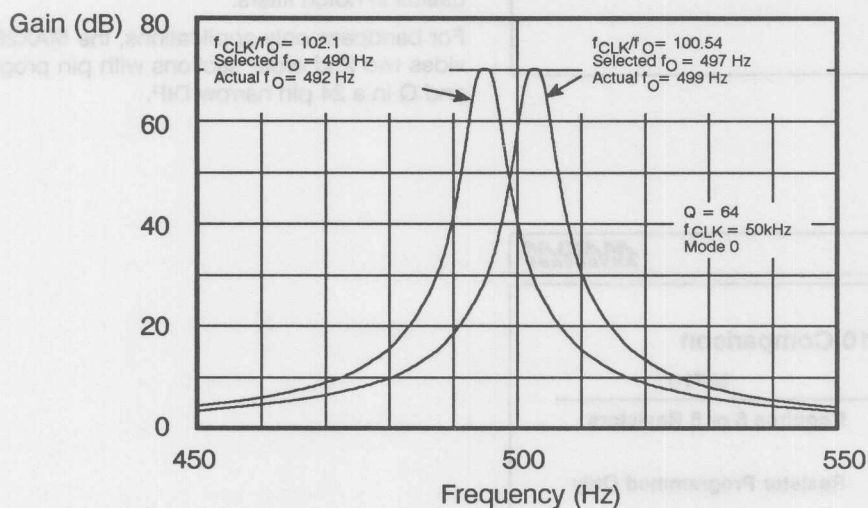
SECOND ORDER BANDPASS, Q = 0.5 to 64



This set of curves shows the measured frequency response of one half of a MAX262. It is set as a 500Hz bandpass filter with Q values

ranging from 0.5 to 64. The Q is set via 7 programming pins. The input clock is 50kHz and f_{CLK}/f_C is set to 100 (Actually 100.54)

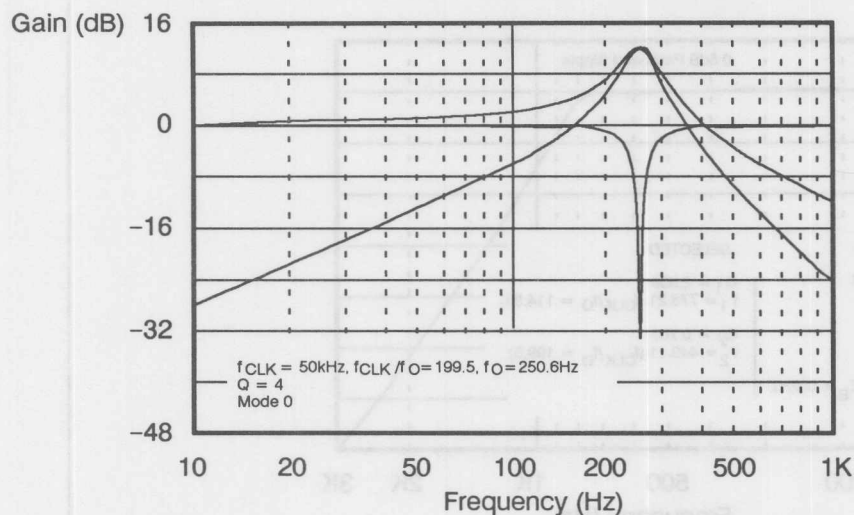
f_O STEP SIZE



The two curves show the minimum step size and center frequency error for a programmed change in bandpass center frequency. The input clock is 50kHz and f_{CLK}/f_C is changed from 100.54 to 102.1.

The expected (selected) center frequencies are 497Hz and 490Hz. The actual frequencies are 499Hz and 492Hz, an error of less than 0.5%. The operating mode for the filter is Mode 0.

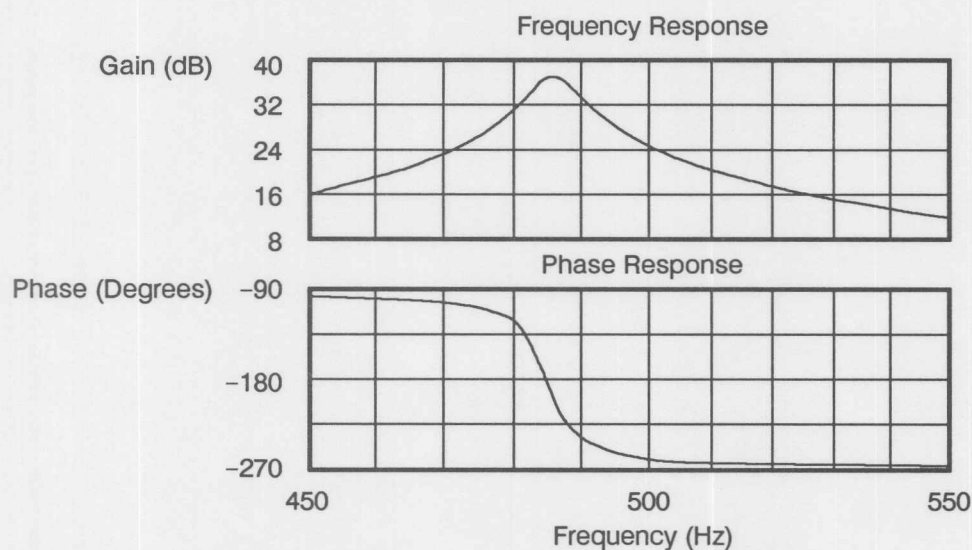
SECOND ORDER BANDPASS, LOWPASS, NOTCH



These are the measured lowpass, bandpass, and notch responses of a MAX262 set for an f_O of 250Hz (Mode 0).

f_{CLK} is 50kHz and f_{CLK}/f_O is set at 199.5. The Q is programmed to 4.

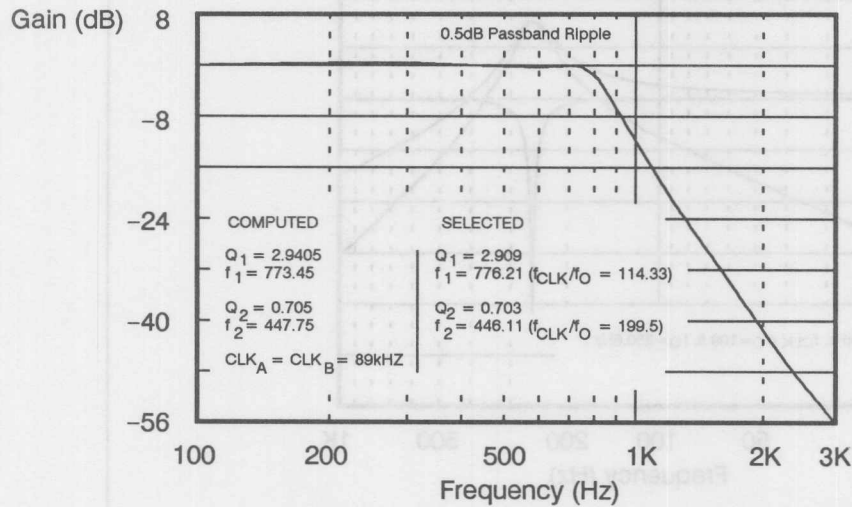
SECOND ORDER BANDPASS



The measured gain and phase for a 2nd order bandpass are shown. The phase response goes from -90 to -270 degrees because each

2nd order block in the MAX260 family is inverting. The phase shift at the center frequency is then -180 degrees rather than 0.

750Hz FOURTH ORDER CHEBYSHEV LOWPASS



The curve is the measured response of a 750Hz 4th order Chebyshev lowpass with 0.5dB passband ripple. The desired f_o and Q for each 2nd order section of such a filter are calculated as:

$$f_1 = 773.45 \quad Q_1 = 2.9405$$

$$f_2 = 447.75 \quad Q_2 = 0.705$$

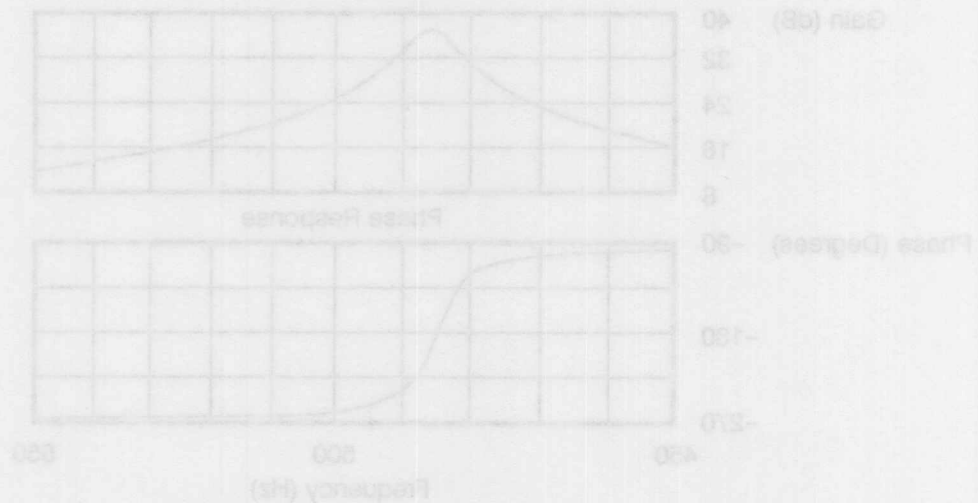
As can be seen, the Q and f_o resolution of the MAX260 family allow high order filters to be programmed without difficulty.

The actual values that can be programmed with a MAX260 are:

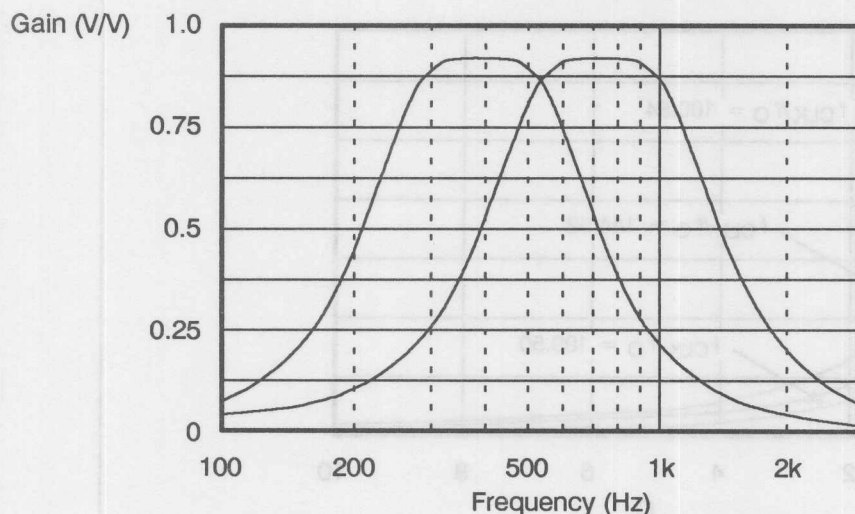
$$f_1 = 776.21 (f_{CLK}/f_O = 114.33) \quad Q_1 = 2.909$$

$$f_2 = 446.11 (f_{CLK}/f_O = 199.5) \quad Q_2 = 0.703$$

$$f_{CLK} = 89KHz$$



FOURTH ORDER BUTTERWORTH BANDPASS



These 4th order bandpass response curves are formed by cascading two 2nd order bandpass filters that have identical Qs but different f_0 frequencies. In this example, different clock frequencies are used for each half of the filter so that the passband can be shifted with a single programmed code change.

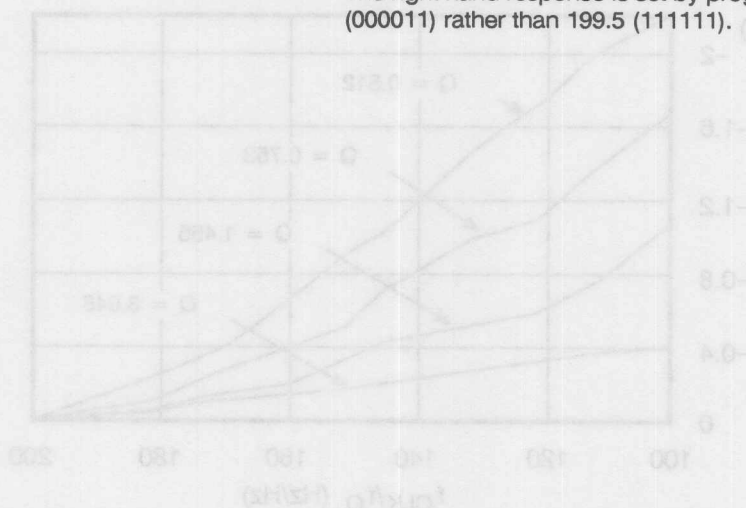
For the left hand response curve:

$$f_1 = 263.7 \quad Q_1 = 1.45 \quad f_{CLK1} = 52.5\text{kHz}$$

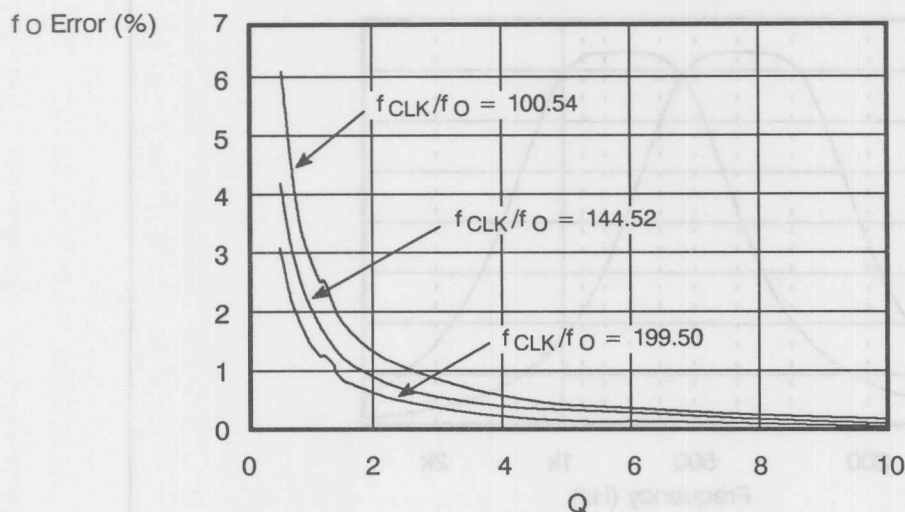
$$f_2 = 568.6 \quad Q_2 = 1.45 \quad f_{CLK2} = 114\text{kHz}$$

$$(f_{CLK}/f_0 \text{ is set to } 199.5 \text{ for } f_1 \text{ and } f_2)$$

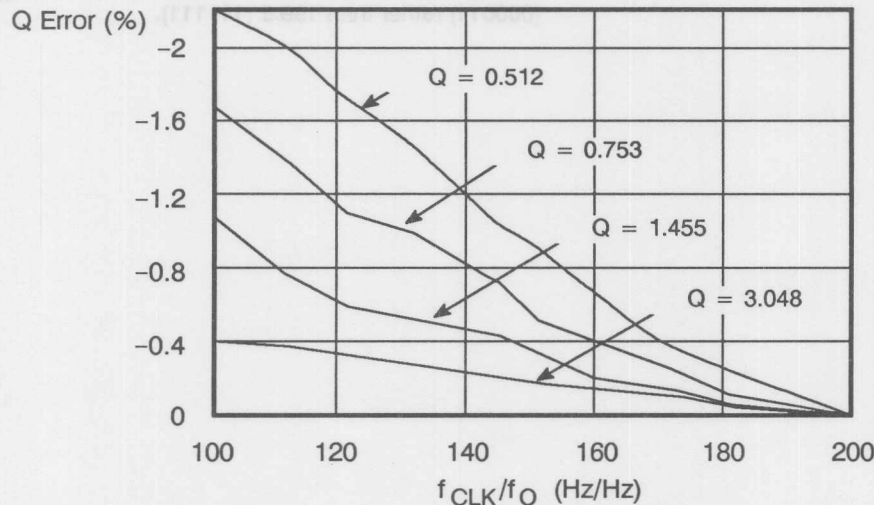
The right hand response is set by programming f_{CLK}/f_0 to 105.24 (000011) rather than 199.5 (111111).



CENTER FREQUENCY ERROR vs Q SETTING



Q ERROR vs f_{CLK}/f_0



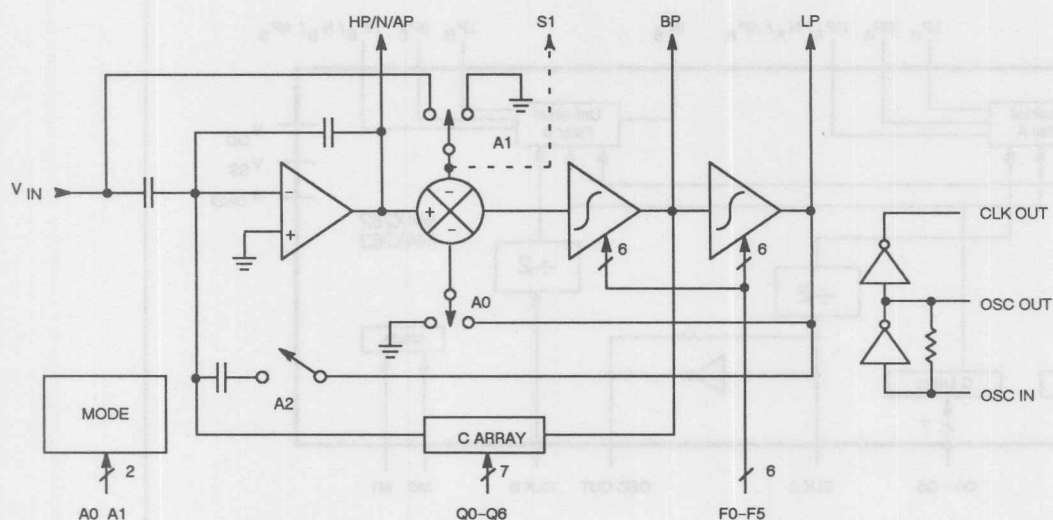
The switched capacitor filter implementation has a slightly different frequency response from what is obtained from an equivalent passive RLC filter. The difference is caused by sampled data effects associated with the finite sampling frequency. The magnitude of this effect is shown in these two graphs.

A sampled data transfer function is usually written with Z transforms. The frequency response is obtained by setting:

$$z = e^{j\omega T} = 1 + j\omega T + (j\omega T)^2/2 + \text{higher order terms}$$

where T is the sampling period. For high sampling rates compared to the frequency of interest, ωT is small and the expression can be approximated by the first two terms. If ωT is large, the higher order terms cause slight differences in filter response compared to active continuous filters. In addition, the second order effects caused by sampling increase as Q decreases (see graph). Because of this it is difficult to precisely set Q and f_0 in a sampled filter, especially when low values are set for Q. Fortunately, this built in error is predictable and can be compensated by adjusting the selected f_0 and Q values.

MAX260 BLOCK DIAGRAM

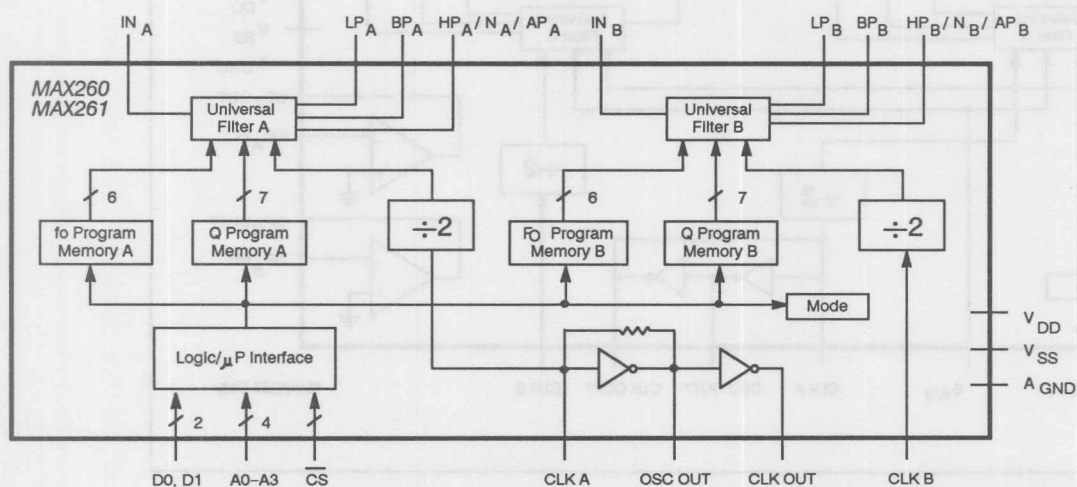


Each 2nd order building block in the MAX260-267 contain two switched capacitor integrators, an input summing amplifier, clock circuitry, and programming inputs for operating mode, Q, and f_0 .

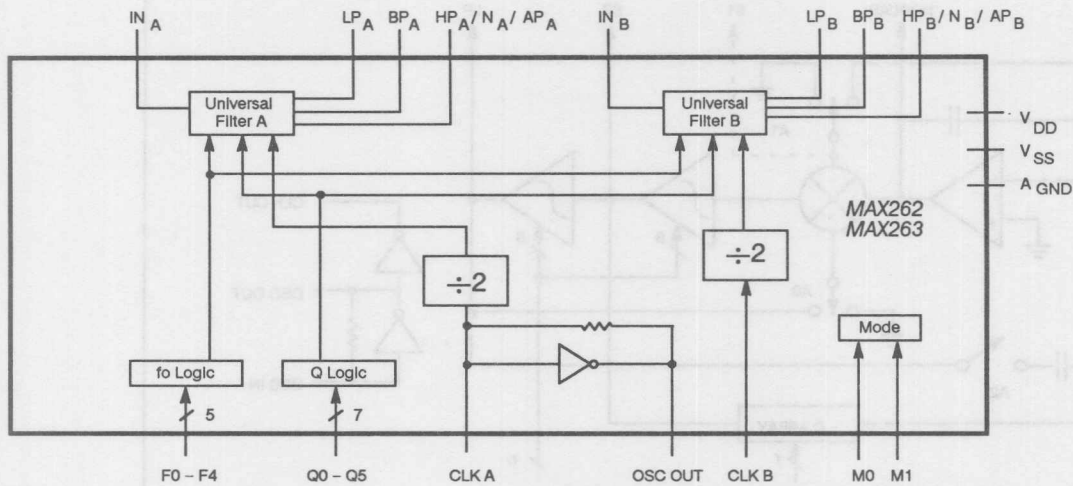
The outputs of the summing amp and integrators provide the filter's highpass/notch/allpass (HP/N/AP), bandpass (BP), and lowpass (LP) outputs.

**MAXIMUM
ADVANTAGE**

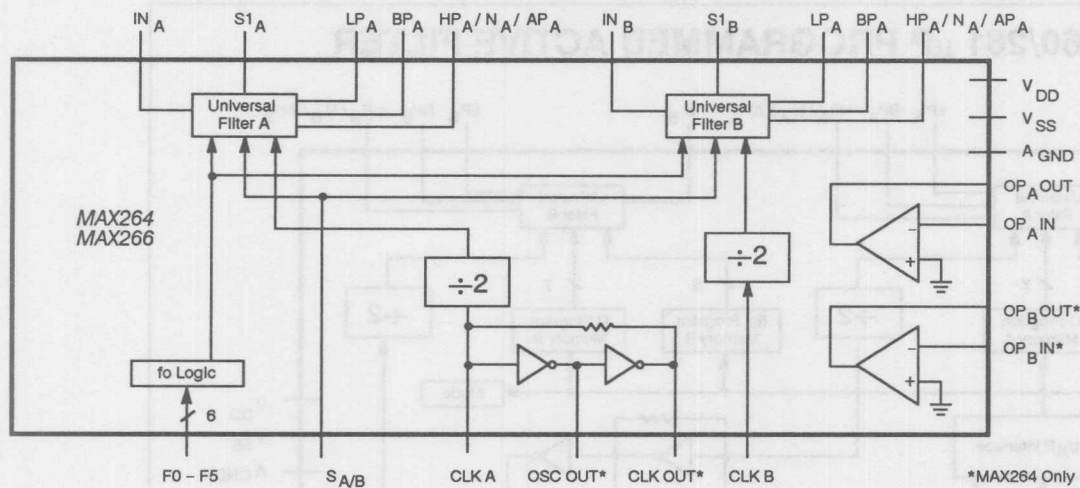
MAX260/261 μ P PROGRAMMED ACTIVE FILTER



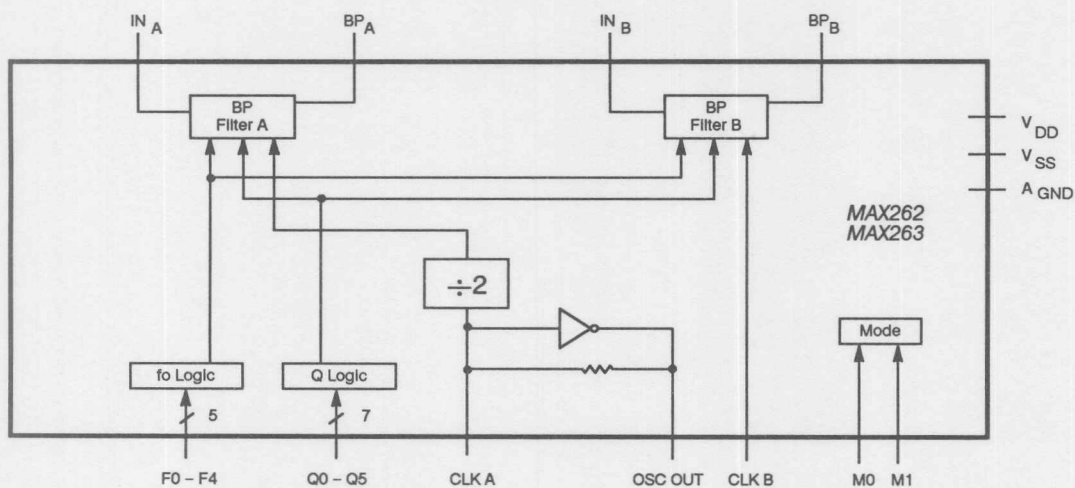
MAX262/263 PIN PROGRAMMED ACTIVE FILTER



MAX264/266 RESISTOR/PIN PROGRAMMED ACTIVE FILTER



MAX268/269 PIN PROGRAMMED BANDPASS FILTER



With almost all manufacturers now manufacturing in the SO package for devices below 20 pins, there are a variety of package styles and dimensions to choose from. This section of the document provides information on the various SO packages and their dimensions. Maxim offers SO for all monolithic devices up to and including 28 pins, and the PLCC package for all 48 pin devices. Maxim can also supply other packages such as the 44 pin flatpack, and it is available on selected devices.

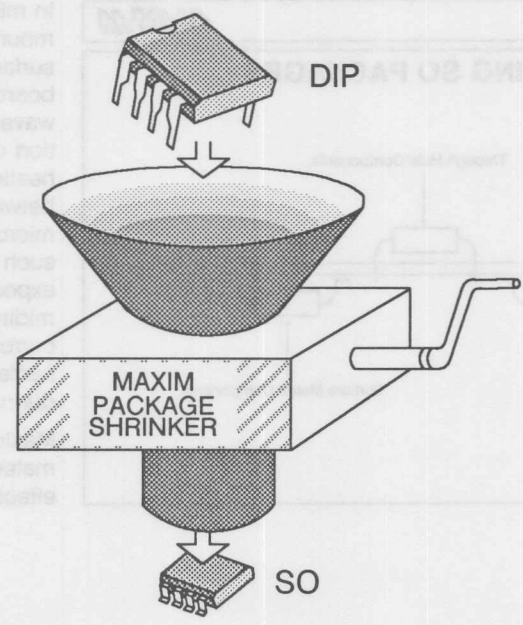
SURFACE-MOUNT PACKAGE OPTIONS					
Package Height	Lead Configuration	Solder Joint	Availability	Industry	Applications
0.1"	0.1"	0.1"	Good	Good	Good
0.15"	0.15"	0.15"	Good	Good	Good
0.2"	0.2"	0.2"	Good	Good	Good
0.25"	0.25"	0.25"	Good	Good	Good
0.3"	0.3"	0.3"	Good	Good	Good



SURFACE MOUNTED PACKAGING

In mixed boards where both through hole and surface mount components are used, it is common to mount the surface mount components on the solder side of the board. This exposes components to a 250°C solder wave. Unless very careful attention is paid to the selection of packaging material and processes, the rapid heating to 250°C causes warpage at the interface between the copper leadframe and the plastic. These warpage stresses allow moisture and corrosive contaminants to enter the package. Subsequent exposure of the board and the components to high humidity can cause premature failure of the circuit due to corrosion of the aluminum on the die by the contaminants that entered the package through the microcracks in the plastic.

Wave soldering has been conducted extensive testing and package material and processing development to reduce these effects to a minimum.



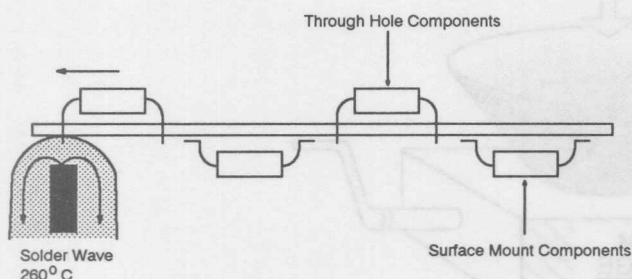
- ## ASSEMBLY PROCEDURES
- ### STRESS PLASTIC PACKAGES
- Thermal shock during solder immersion up to 250°C produces microcracks in "high stress" epoxy SO packages (14 and 16-pin 0.75" wide SO).
 - Corrosive contaminants enter package (flux/solder paste).
 - Subsequent moisture ingress causes metalization causing premature failure.

SURFACE-MOUNT PACKAGE OPTIONS

	Small Outline	PLCC	LCC	Flat-Pack
Package Height	0.1"	0.2"	0.1"	0.1"
Lead Compliance	Excellent	Poor	None	Excellent
Solder-Joint Visibility	Good	Poor	Poor	Good
Industry Acceptance (New Designs)	Excellent	Good	Good	Poor

While almost all manufacturers have standardized on the SO package for devices below 28 pins, there are a variety of package styles and dimensions for devices of 28 leads and above. Maxim offers SO for all monolithic devices up to and including 28 leads, and the PLCC package for all 40 pin devices. Maxim can also supply other packages such as the 44 pin flatpack, and it is available on selected devices.

WAVE SOLDERING SO PACKAGES



In mixed boards where both through hole and surface mount components are used, it is common to mount the surface mount components on the solder side of the board. This exposes components to a 260°C solder wave. Unless very careful attention is paid to the selection of packaging material and procedures, the rapid heating to 260°C creates microcracks at the interface between the copper leadframe and the plastic. These microcracks allow moisture and corrosive contaminants such as solder flux to enter the package. Subsequent exposure of the board and its components to high humidity can cause premature failure of the circuit due to corrosion of the aluminum on the die by the contaminants that entered the package through the microcracks during wave soldering.

Maxim has conducted extensive testing and package material and processing development to reduce these effects to a minimum.

ASSEMBLY PROCEDURES STRESS PLASTIC PACKAGES

- Thermal Shock during solder immersion, up to 260°C, produces Microcracks in "High Stress" epoxy SO packages (8, 14 and 16-pin, 0.15" wide SO)
- Corrosive contaminants enter package (flux/solder paste)
- Subsequent moisture ingress corrodes metalization causing premature failure

To simulate the stress on wave soldering, Maxim qualifies its SO packages by a sequence of three 260°C solder immersions followed by a 96 hour pressure cooker test under the standard conditions of 121°C, 15PSI, 100% relative humidity.

MAXIM SOLVES THE PROBLEM

Low Stress Epoxy Withstands PCT After Thermal Shock

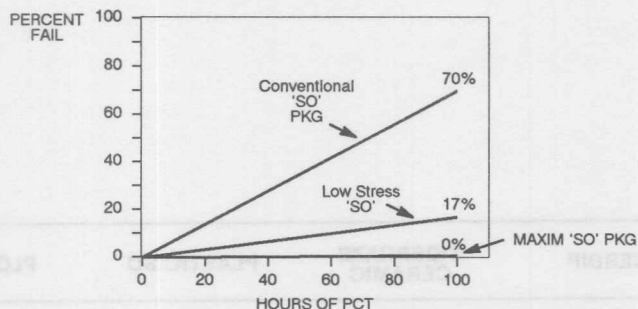
1. 7 Day, 30°C 85%RH Preconditioning
2. 10 Second, 260°C Solder Immersion
3. Repeat Solder Immersion Two More Times
4. 96Hr Pressure Cooker Test @121°C, 15PSI

The data is for the worst case device, an 8 pin 0.150" wide plastic SO. Most manufacturer's packages use "high stress" epoxy, and will have a failure rate of about 70% after 96 hours if the SO package is subjected to a solder immersion test prior to the pressure cooker test.

The second generation "low stress" material has an unacceptable failure rate of 17%, and the material used by most Japanese manufacturers still shows a 1.5% failure rate.

In comparison, the 4th generation material and process techniques for Maxim SO packages since August 1986 typically has 0 failures after 96 hours of pressure cooker test, even when they are first subjected to the 260°C solder immersion test.

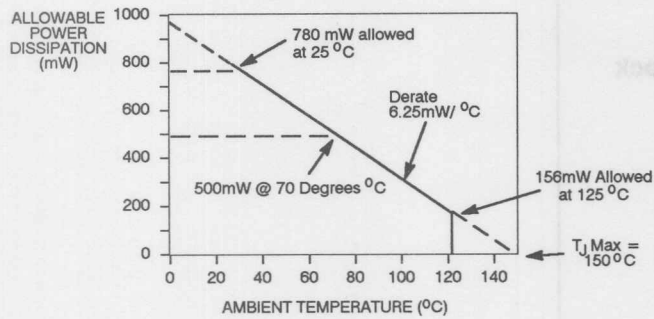
SO RELIABILITY COMPARISON (PCT FOLLOWING SOLDER IMMERSION TEST)



SO PACKAGE FAILURES AFTER STRESS TEST

EPOXY MATERIAL	FAILURES AFTER SOLDER IMMERSION & PCT	
HIGH-STRESS	22/25	TOTAL 109/156, or 70%
	41/41	
	22/45	
	24/45	
POPULAR LOW-STRESS	5/25	TOTAL 27/158, or 17%
	5/43	
	6/45	
	11/45	
JAPANESE SO	0/25	TOTAL 1/69, or 1.5%
	1/44	
MAXIM LOW-STRESS SO	0/25	TOTAL 0/108, or 0%
	0/38	
	0/45	

MAXIMUM ALLOWABLE POWER DISSIPATION vs TEMPERATURE - 8 PIN PLASTIC DIP



This graph shows the maximum allowable power dissipation for an 8 pin plastic DIP with an alloy 42 or Kovar leadframe. On most data sheets the maximum allowable power dissipation is stated in the Absolute Maximum Ratings section of the specification table. The power dissipation specification generally comes in two parts: a maximum power dissipation for all temperatures below a somewhat arbitrarily chosen temperature; and a derating factor for calculating the allowable power dissipation for higher temperatures. The absolute maximum rating on power dissipation that corresponds to the graph is "500mW up to 70°C, derate 6.25mW/°C above 70°C". An alternative spec would be "780mW up to 25°C, derate 6.25mW/°C above 25°C".

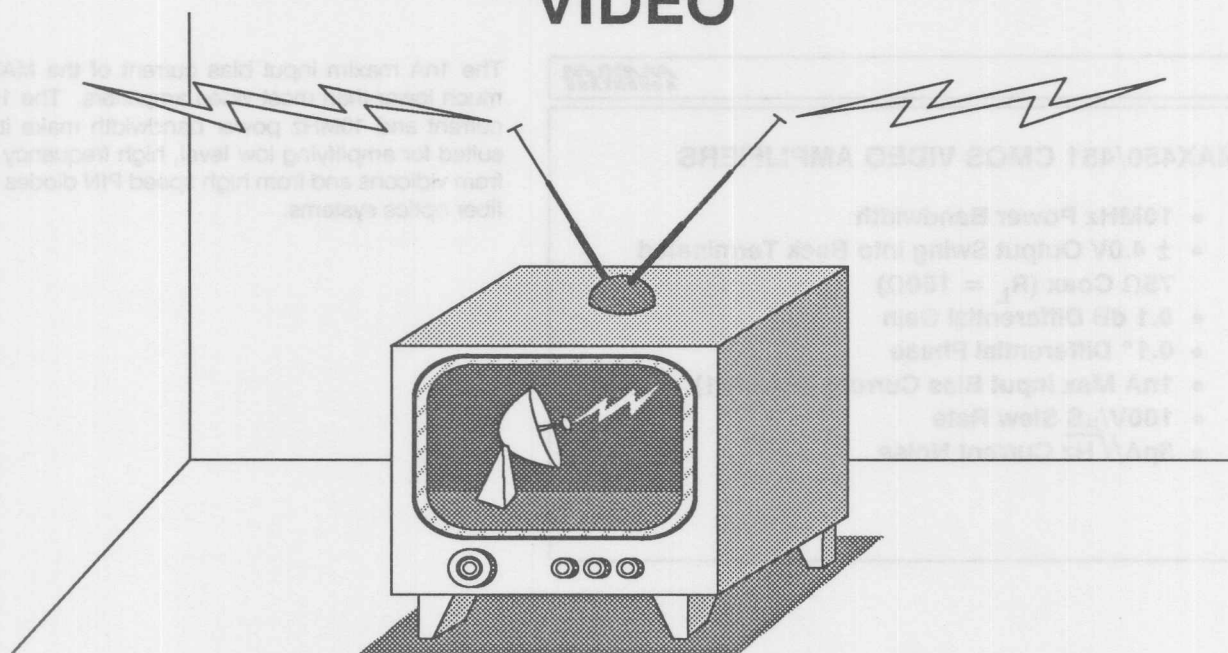
The purpose of the power dissipation limit is ensure that the die temperature does not exceed its maximum allowable temperature, in this case, 150°C. The slope of the allowable power dissipation vs. ambient temperature is the reciprocal of the θ_{JA} . As shown in the table, the θ_{JA} for an 8 pin plastic DIP with alloy 42 leadframe is 160°C/W. The reciprocal is 6.25mW/°C, which is the derating factor of this device.

Thermal Resistance of Maxim Packages

(θ_{JA}/θ_{JC} in °C/Watt)

LEADS	PLASTIC WITH ALLOY 42 LEAD FRAME	PLASTIC WITH COPPER LEAD FRAME	CERDIP	SIDE BRAZE CERAMIC	PLASTIC SO	PLCC
8	160/75	120/70	125/55	120/50	170/80	
14	140/70	100/60	105/50	100/45	115/60	
16	135/65	100/60	100/50	95/45	110/60	
18	130/60	95/60	95/45	90/40	105/60	
20	125/60	90/50	90/40	85/35	100/60	
24 (300 mil)	120/60	85/50	80/40	—	90/50	
24 (600 mil)	110/50	75/45	55/20	50/15	85/45	
28	110/50	70/45	55/20	50/15	80/45	100/45
40	100/45	60/40	45/20	40/15		
44						80/40

VIDEO



Video Signal Products and Applications

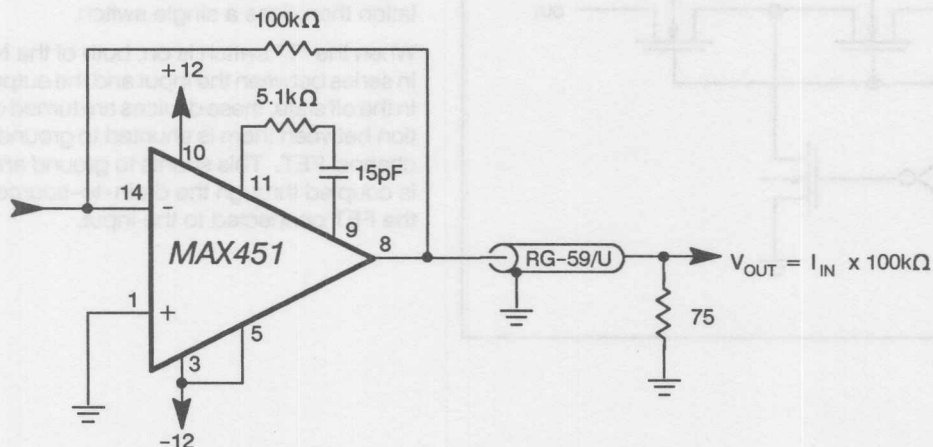
- MAX450/451 Video Amplifier
- Vidicon Amplifier
- MAX452-455 Video Multiplexer/Amplifier
- Video Switches and Multiplexers
- Coax Powered Video Multiplexer

MAX450/451 CMOS VIDEO AMPLIFIERS

- 10MHz Power Bandwidth
- $\pm 4.0\text{V}$ Output Swing into Back Terminated 75Ω Coax ($R_L = 150\Omega$)
- 0.1 dB Differential Gain
- 0.1° Differential Phase
- 1nA Max Input Bias Current (MAX451)
- $100\text{V}/\mu\text{S}$ Slew Rate
- $5\text{pA}/\sqrt{\text{Hz}}$ Current Noise

The 1nA maxim input bias current of the MAX451 is much lower than most video amplifiers. The 1nA bias current and 10MHz power bandwidth make it is well suited for amplifying low level, high frequency signals from vidicons and from high speed PIN diodes used in fiber optics systems.

VIDEO



Vidicon or PIN Diode Transimpedance Amplifier
(Bandwidth = 5MHz)

For the widest bandwidth, the stray capacitance across the 100kΩ feedback resistor must be minimized. One way to minimize the stray capacitance is to use five 20kΩ resistors in series instead of

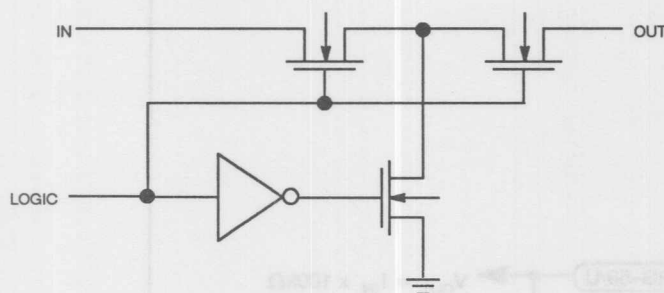
just one 100kΩ resistor. Just 1pF of capacitance across the 100kΩ resistor will cause the system to have a pole at 1.6MHz.

MAX452/3/4/5 VIDEO MUX/AMPLIFIERS

- 50MHz Gain Bandwidth
- Onboard Multiplexer
 - 8 Channels -- MAX455
 - 4 Channels -- MAX454
 - 2 Channels -- MAX453
- High Impedance Inputs
 - 10pA Typical I_{BIAS}
 - 9pF Input Capacitance
- 2pF On-State-Off-State Capacitance Change
- Drives 75Ω Coax $\pm 1V$
- Unity Gain Stable
 - No External Compensation Needed

The MAX452 50MHz gain bandwidth video amplifier is made in a silicon gate CMOS process. This enabled Maxim to add an onboard multiplexer to make the MAX453, MAX454 and MAX455, which have a 2, 4, and 8 channel input multiplexer respectively. Since the multiplexer is onboard, the output of the mux need only drive the very small input capacitance of the CMOS amplifier. Because of this, the analog switch impedance can be as high as 2kΩ. This, in turn keeps the input capacitance of each channel down to 7pF in the off state and 9pF when selected.

VIDEO "T" SWITCH



A "T" configuration switch is used in three Maxim product families: the IH5341/5352 dual and quad video switches, the MAX310/311 multiplexer family, and the MAX453/4/5 video mux/amplifier family. The "T" switch is used because it has lower feedthrough capacitance, and therefore, higher off isolation and inter-channel isolation than does a single switch.

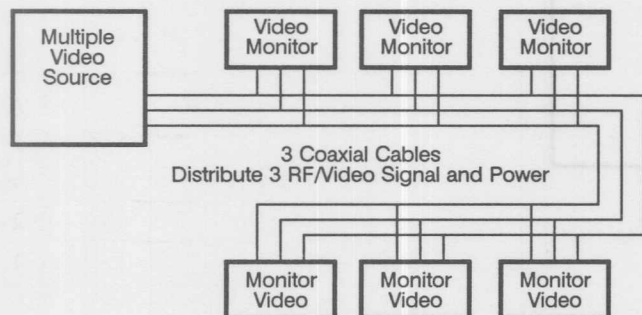
When the "T" switch is on, both of the N-channel FETs in series between the input and the output are turned on. In the off state, these devices are turned off, and the junction between them is shunted to ground via the third N-channel FET. This shunts to ground any AC signal that is coupled through the drain-to-source capacitance of the FET connected to the input.

Video Switch/Mux Features

Video Switches and Multiplexers

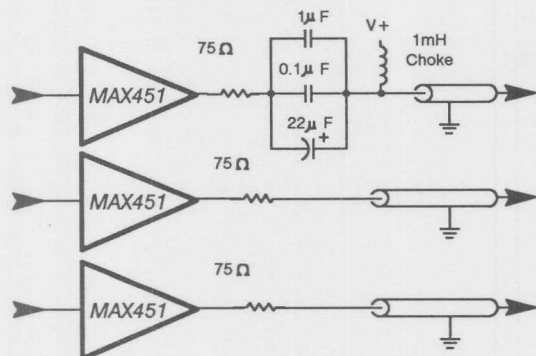
- Dual (IH5341) and Quad (IH5352) SPST Switches
- 1 of 8 Multiplexer --- MAX310
- 2 of 4 Multiplexer --- MAX311
- 70dB Minimum Off Channel Isolation at 10MHz (IH5341/5352)
- $\pm 5V$ to $\pm 15V$ Operation
- 75Ω $r_{DS(on)}$ Matches Coaxial Cable Impedance (IH5341/5352)
- 3dB Bandwidth of 100MHz Minimum

VIDEO DISTRIBUTION SYSTEM



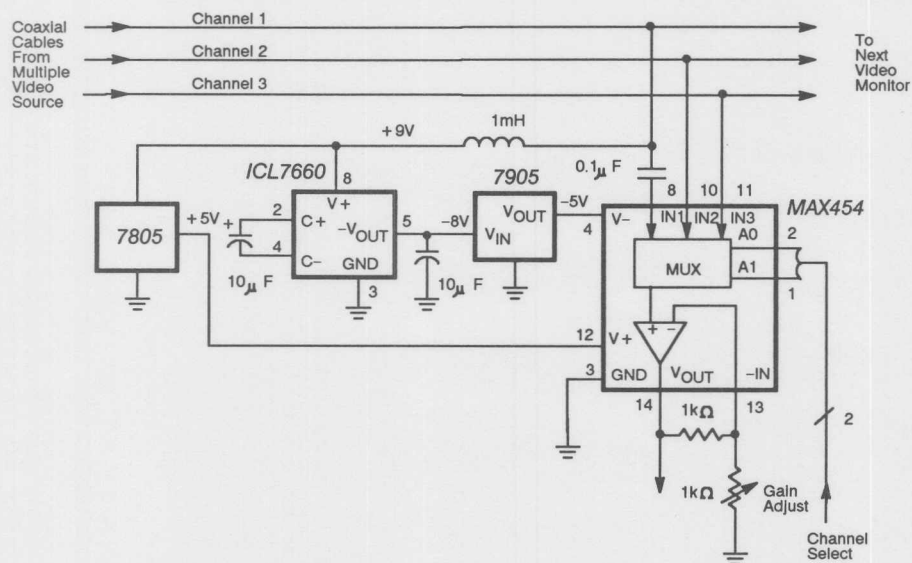
This is a simplified block diagram of a typical video distribution system where three video signals are distributed by coaxial cable to 6 terminals. Each terminal can select any of the 3 video signals. The potential problem in this type of system is that if power is turned off to one of the terminals, it may short out one or more of the video signals. This can be avoided if both power and video is distributed using the same coaxial cable.

VIDEO AND POWER SOURCE



+9V can be coupled into the center conductor of one of the coaxial cables by using a 1mH choke to isolate the +9V supply from the AC coupled video signal. The MAX450 video amplifier can drive 75Ω back terminated lines such as these to $\pm 4V$.

VIDEO MULTIPLEXER POWERED FROM COAX



At the terminal, the DC voltage is recovered through a 1mH choke, inverted to -8V by an ICL7660 and regulated to $\pm 5V$. MAX454 video amplifier with multiplexer combines both the video switching

for selecting one of the 3 signal sources, and an amplifier for adjusting the video output voltage level.

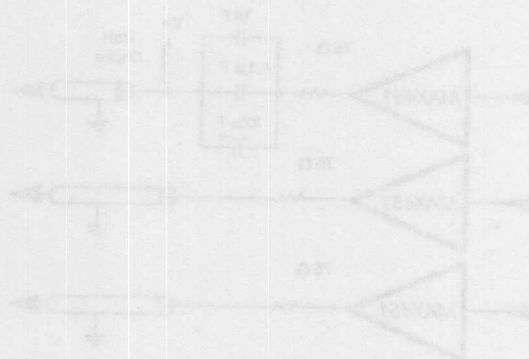
Video Buffers and Op Amps

PART NUMBER	TYPE	BANDWIDTH (MHz)	SLEW RATE (V/ μ s)	OUTPUT CURRENT	FEATURES
BB3553	Video Buffer	300	6000	200mA	Drive 50Ω loads
BB3554	Video Op Amp	1700	1000	100mA	High Gain
LH0033	Video Buffer	100	1500	100mA	Drives 75Ω loads
LH0063	Video Buffer	300	6000	200mA	Drives 50Ω loads
MAX450	Video Op Amp	10	70	30mA	Drives 75Ω loads
MAX451	Video OP Amp	10	70	30mA	1nA max I_{BIAS}
MAX460	Video Buffer	100	1500	100mA	Low I_B and C_{IN}

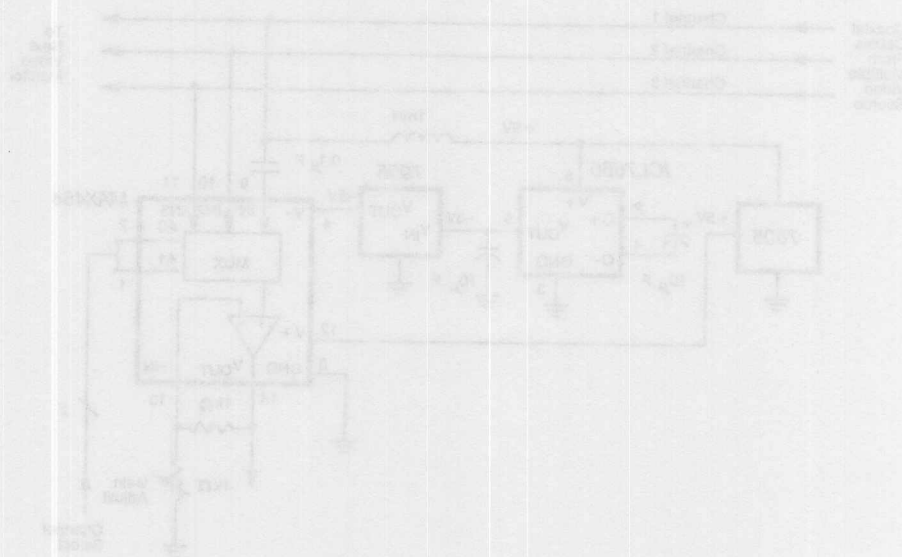
NOTES

±5V can be coupled into the center conductor of one of the coaxial cables by using a 1mm choke to isolate the ±5V supply from the AC coupled video signal. The MAX420 video amplifier can drive 75Ω back terminated lines such as these to ±5V.

VIDEO AND POWER SOURCE



VIDEO MULTIPLEXER POWERED FROM COAX



At the terminal, the DC voltage is recovered through a high choke. The video is recovered through a high choke. The video is recovered through a high choke. The video is recovered through a high choke.

for selecting one of the signal sources, and an amplifier for each of the video output voltage level.

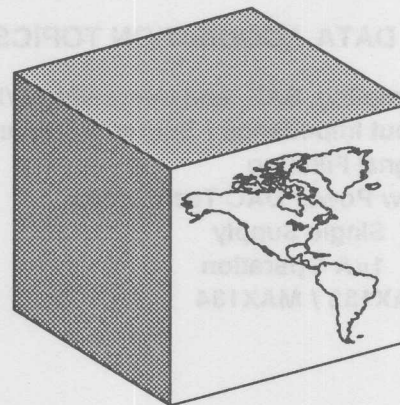
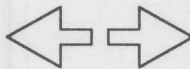
Video Buffers and Op Amps

PART NUMBER	TYPE	BANDWIDTH (MHz)	SIGNAL RATE (Vrms)	OUTPUT CURRENT	FEATURES
MAX420	Video Buffer	200	8000	200mA	Drive 50Ω loads
MAX421	Video Op Amp	100	4000	100mA	High Gain
MAX422	Video Buffer	100	1800	100mA	Drive 75Ω loads
MAX423	Video Buffer	80	8000	200mA	Drive 50Ω loads
MAX424	Video Op Amp	10	10	20mA	Drive 75Ω loads
MAX425	Video Op Amp	10	10	20mA	7mA max
MAX426	Video Buffer	100	1200	100mA	Low 1/2 max Op

PART NUMBER	RESOLUTION	INTERNAL LINEARITY	CONVERSION TIME	SAMPLE VOLTAGE RANGE	INPUT RANGE	FEATURES
MAX100	8 bits	$\pm 1/2 \text{LSB}$	1.0 μs	$\pm 5\text{V}$	$\pm 5\text{V}$	Internal Reference
MAX104	8 bits ch	$\pm 1/2 \text{LSB}$	2.0 μs	$\pm 5\text{V}$	$\pm 5\text{V}$	Internal Reference
MAX105	8 bits ch	$\pm 1/2 \text{LSB}$	2.0 μs	$\pm 5\text{V}$	$\pm 5\text{V}$	Internal Reference
MAX107	8 bits	$\pm 1/2 \text{LSB}$	4.0 μs	$\pm 5\text{V}$	$\pm 10\text{V}$	Full A/D 10V
MAX110	8 bits ch	$\pm 1/2 \text{LSB}$	20 μs	$\pm 5\text{V}$	$\pm 10\text{V}$	Full A/D 10V
AD7810	10 bits	$\pm 1/2 \text{LSB}$	3 μs	$\pm 10\text{V}$	$\pm 10\text{V}$	Onboard Reference
AD7820	8 bits	$\pm 1/2 \text{LSB}$	1.0 μs	$\pm 5\text{V}$	$\pm 5\text{V}$	Half-Power
AD7825	12 bits	$\pm 1/2 \text{LSB}$	2.0 μs	$\pm 5\text{V}$	$\pm 5\text{V}$	Internal Reference
AD7828	8 bits	$\pm 1/2 \text{LSB}$	1.0 μs	$\pm 5\text{V}$	$\pm 10\text{V}$	Autozero > 10V
AD7829	8 bits	$\pm 1/2 \text{LSB}$	50 μs	$\pm 5\text{V}$	$\pm 10\text{V}$	8-channels

**MAXIM
ADVANTAGE**

ANALOG DATA CONVERSION



Data Acquisition A/D Converter Selector

PART NUMBER	RESOLUTION	INTEGRAL LINEARITY	CONVERSION TIME	SUPPLY VOLTAGE	INPUT RANGE	FEATURES
MAX150	8 Bits	1/2 LSB	1.34μs	+5V	+5V	Internal Reference
MAX154	8 bits/4 ch	1/2 LSB	2.0μs	+5V	+5V	Internal Reference
MAX158	8 bits/8 ch	1/2 LSB	2.0μs	+5V	+5V	Internal Reference
MAX160	8 bits	1/2 LSB	4μs	+5V	±15V	Fast AD7574
MAX161	8 bits/8 ch	1/2 LSB	20μs	+5V	±15V	Fast AD7581
AD578	12 bits	1/2 LSB	3μs	±15V	±10V	Onboard Reference
ADC0820	8 bits	1/2 LSB	1.4μs	+5V	+5V	Half-Flash
AD7572	12 bits	1/2 LSB	5.0μs	+5/-15V	+5V	Internal Reference
AD7574	8 bits	1/2 LSB	15μs	+5V	±15V	Analog $V_{IN} > V_{SUPP}$
AD7581	8 bits	1/2 LSB	66.6μs	+5V	±15V	8 byte RAM
AD7820	8 bits	1/2 LSB	1.34μs	+5V	+5V	Half-Flash
AD7824	8 bits/4 ch	1/2 LSB	2.0μs	+5V	+5V	4 Channels
AD7828	8 bits/4 ch	1/2 LSB	2.0μs	+5V	+5V	8 Channels



DATA ACQUISITION TOPICS

Sampling, SAR, and Integrating A/Ds
 Input Impedance / Source Resistance
 Signal Filtering
 Low Power DAC Techniques
 Single Supply
 1μA Operation
 MAX133 / MAX134

CMOS analog-to-digital converters improve conversion performance in data acquisition systems while reducing costs and simplifying designs. Many new devices employ conversion techniques that are new or modified versions of classic architectures. This results in slightly different input behavior on some converters, which in most cases is transparent. Nevertheless, understanding these differences helps keep analog "bugs" at bay.

Although CMOS digital-to-analog converters have always been thought of as low power devices, it is not widely known just how low they can be.

The MAX133/134 is a new uP interfactable integrating A/D. It is used in both digital readout and system applications. Internal resolution is $\pm 40,000$ counts with features such as auto zero calibration and input multiplexer.

DATA ACQUISITION A/D CONVERTERS (Data output as opposed to display output)

	BITS	SPEED	CHANNELS	INPUT TYPE
MAX150	8	1.34us	1	Sampling
MAX154/158	8	2.0us	4/8	Sampling
MAX160	8	5us	1	SAR
MAX161	8	20us	8	SAR
MAX133/134	+/-15*	50ms	7	Integrating
AD7574	8	15us	1	SAR
AD7581	8	67us	8	SAR
AD7820	8	1.34us	1	Sampling
AD7824/28	8	2.0us	4/8	Sampling
ADC0820	8	1.4us	1	Sampling
AD7572	12	5us	1	SAR
ICL7109	+/-12	33ms	1	Integrating
ICL7135	+/-14	330ms	1	Integrating

* The MAX133/134 has +/-40,000 count internal resolution

This is a list of Maxim's current data acquisition A/D products, i.e. those devices which convert an analog input (or inputs) to a digital data output as opposed to a display output. The last item in each listing describes the converter type: SAMPLING (flash or half-flash), SAR (successive approximation), or INTEGRATING.

High Speed 8 Bit A/Ds with Reference MAX150/154/158

- Fast Conversion Time
MAX150 – 1.34us
MAX154/8 – 2.5us
- 1, 4, or 8 Analog Channels
- Internal 2.5V Reference
- Built In Track/Hold Function
- Single +5V Operation
- Easy Microprocessor Interface
- Compatible with AD7820/24/28

MAX160 uP Compatible 8 Bit A/D

- Pin Compatible with AD7574 but
Almost 4x Faster: 4 vs. 15us
- No External Clock Required
- Single +5V Supply
- 25mW Max Power Consumption
- Three Interface Modes Simplify
Microprocessor Connections

MAX161 8 Channel Data Acquisition System

- Pin Compatible with AD7581 but
Over 3x Faster: 20 vs. 67us
- 8 bit Accuracy and Resolution
- Direct Z80, 8085, 6800 Interface
- Interleaved DMA Operation
- On-Chip 8x8 Dual Port RAM
- Single +5V Supply

INPUT SOURCE RESISTANCE

Sampling

R_{IN} = high, but transient input current must settle.

$R_S < 500\Omega$ for $< 1/2$ LSB error

SAR

$R_{IN} = 20k\Omega$ to summing node of comparator

$R_S = 100\Omega$: approx. 0.5% error

Integrating

R_{IN} = very high

$R_S = \text{up to } 100M\Omega$: < 1 LSB error

The input behavior of an A/D varies with converter type. How an A/D converter appears to an analog signal is of prime importance, especially in high speed and/or high resolution stems. CMOS converters are extremely trouble-free in this regard, but an understanding of the basic operations involved will help keep analog bugs at bay.

In general CMOS A/Ds have high input impedance and so can measure signals from a variety of sources without error. The recommended range of input resistance for each converter type is outlined above.

SAMPLING: (MAX150/54/58, AD7820/24/28, ADC0820) These devices have high input resistance but also have input capacitance which must be charged with each conversion. The total input current is a function of conversion rate and input voltage. The allowable source resistance is limited not by the average input current but by the short time available to charge the input capacitance to the signal voltage.

SAR: (MAX160/161, AD7572/74/81): The analog input(s) of most (not all) CMOS SAR A/Ds are modeled as a resistance connected to the current summing node of a comparator. The input current varies during the conversion as the successive approximation search takes place. At the end of the conversion, the input looks like a resistance to ground as the internal DAC balances the input signal.

INTEGRATING: (MAX133, ICL7109/35) Most CMOS integrating converters have extremely high input impedance and low input bias current. Source resistances of up to $100M\Omega$ are frequently accommodated without buffering.

A/D INPUT FILTERING

Sampling

Input filter cap may cause gain error.

SAR

Simple RC filter. C is shunted by R_{IN} .

Integrating

Simple RC input filter. High R_{IN} allows small capacitor values

Noise is always a concern in analog systems. A common approach to noise reduction is low-pass filtering may be as simple as an RC network placed at the A/D input, but this may not always be advisable or even necessary with some A/D types.

SAMPLING: The A/D's input capacitance is charged from the input signal. This produces an input current transient with each conversion. This transient is not a problem as long as it settles before the A/D's comparators make their decision. If a capacitor is connected at the input, the transients are then integrated so that a gain error remains when the comparators are latched.

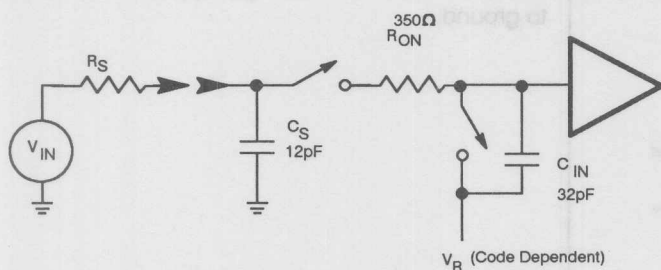
To avoid input error if a filter cap is used, it must be large enough to supply the transient input current without drooping after repetitive conversions. Of course the signal source must be able to keep the capacitor charged as well.

If the input filter is buffered with an op-amp then there are no restrictions on capacitor value since the cap no longer supplies current directly to the A/D input.

SAR: Input filtering on SAR converters generally causes no "hidden" errors. An input capacitor, if used, is shunted by the A/Ds input resistance (typ $10k\Omega$ to $30k\Omega$) so large values of capacitance may be needed for long time constants.

INTEGRATING: Input filtering is easiest with integrating A/Ds. Long time constants are implemented with small capacitors because the analog inputs have such high resistance and low bias current. It's ironic however that integrating A/Ds are least likely to require additional filtering because of their inherent noise rejection.

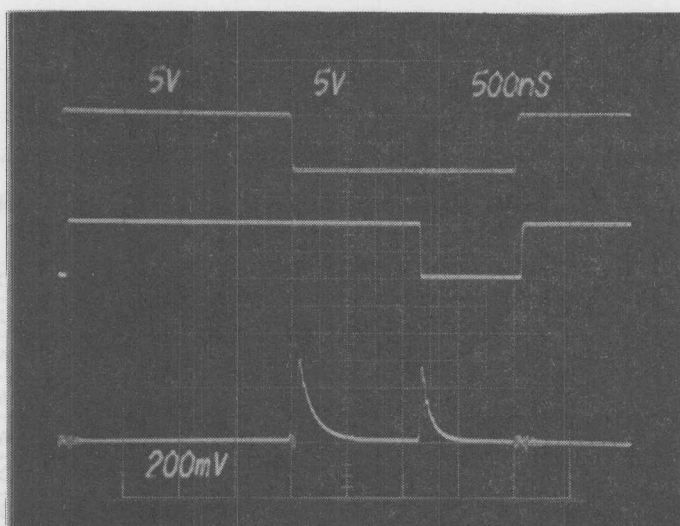
SAMPLING A/D INPUT CURRENT MODEL



The input model for a sampling A/D converter such as the MAX150 is dominated by a switched input capacitance of typically 32pF. This capacitance must be charged with each conversion. V_R is code dependent but is nominally 2.5V. The input current is:

$$I_{IN} = \text{Conv./sec} \times V_{IN} - V_R \times C_{IN}$$

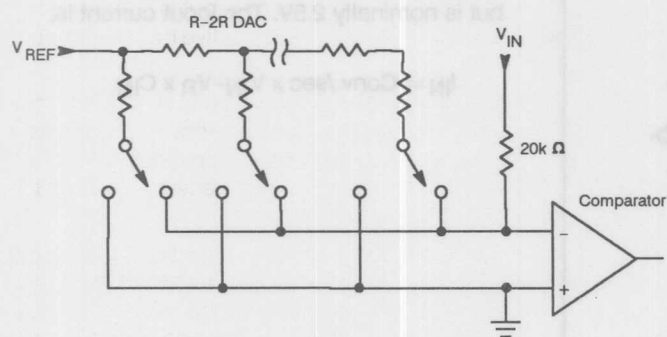
MAX150 INPUT



The photo shows the READ input, INTERRUPT output, and the analog input of a MAX150 running at 250,000 conversion/sec. The input signal is 0V through a 1kΩ source resistance. R_S is larger than recommended to highlight the input transient. Again, no conversion er-

ror is generated by these spikes as long as they are allowed to settle before the internal comparators are latched (approx. 600ns in the READ interface mode).

SAR A/D INPUT MODEL

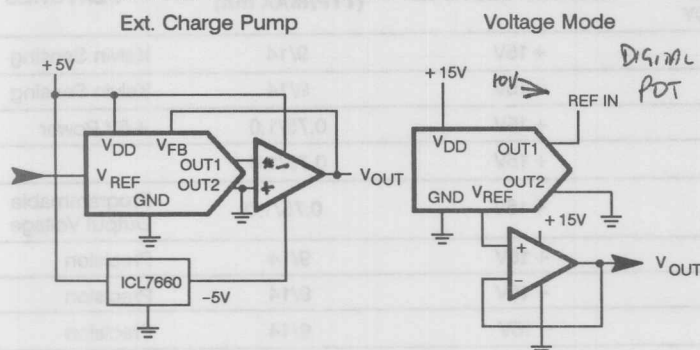


In SAR A/Ds, current from the input (V_{IN}) is compared to an R-2R DAC. The DAC goes through a binary search before settling within an LSB of $-V_{IN}$. The input current changes during the conversion process depending on the specific states of the DAC during the search. At the end of the conversion the input appears as a resistance to ground.

D/A Converters

PART NUMBER	TYPE	RESOLUTION	RELATIVE ACCURACY % F.S	GAIN TEMPCO ppm/°C MAX	SETTLING TIME	OUTPUT	POWER DISSIPATION (mW)
AD565A	Bipolar, w/Ref.	12-bit	0.012 to 0.006	20	250ns max	Current	345
AD566A	Bipolar	12-bit	0.012 to 0.006	20	350ns max	Current	345
AD7224	Multiplying	8-bit	0.2	20	7μs max	Voltage	63
AD7225	Quad, Multiplying	8-bit	0.2	20	5μs max	Voltage	190
AD7226	Quad, Multiplying	8-bit	0.2	20	5μs max	Voltage	190
AD7520	Multiplying	10-bit	0.2 to 0.05	10	500ns typ	Current	20
AD7521	Multiplying	12-bit	0.2 to 0.05	10	500ns typ	Current	20
AD7523	Multiplying, Low Cost	8-bit	0.2 to 0.05	67	150ns typ	Current	1.6
AD7524	Multiplying	8-bit	0.5 to 0.2	40	100ns max	Current	30
AD7528	Dual, Multiplying	8-bit	0.2	35	180ns max	Current	15
AD7530	Multiplying	10-bit	0.2 to 0.05	10	500ns typ	Current	20
AD7531	Multiplying	12-bit	0.2 to 0.05	10	500ns typ	Current	20
AD7533	Multiplying	10-bit	0.2 to 0.05	22	600ns max	Current	30
AD7541	Multiplying	12-bit	0.025 to 0.012	23	1μs max	Current	30
AD7541A	Multiplying	12-bit	0.025 to 0.012	5	600ns typ	Current	7.5
AD7542	Multiplying	12-bit	0.012	5	2μs max	Current	12.5
AD7543	Serial Input	12-bit	0.012	5	2μs max	Current	12.5
AD7545/MP7645	Multiplying	12-bit	0.05 to 0.012	5	2μs max	Current	7.5
AM6012	Bipolar, Multiplying	12-bit	0.025 to 0.012	20	250ns typ	Current	397

SINGLE SUPPLY DAC APPLICATIONS



Although CMOS multiplying DACs operate from one power supply, their common circuit configuration requires a negative supply as well. The minus supply is needed either for the output op-amp, which must swing negatively if the reference input is positive, or for the reference, which must be negative if a positive output is desired.

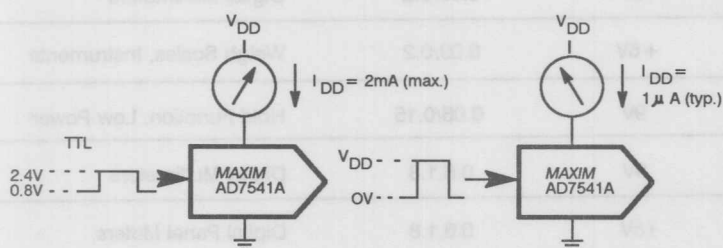
Two alternate DAC connections which DO function from a single positive supply are shown. The left hand circuit uses a standard output op-amp hookup with an ICL7660 voltage inverter to generate the negative power.

The right side shows an inverted connection commonly known as "voltage mode" in which the current outputs (OUT1, OUT2) are used as the reference input, and the reference input (VREF) is used as a voltage output. This circuit does not invert but usually requires an op-amp buffer because the output resistance changes with the DAC setting.

Some newer CMOS DACs do not require special connections for single supply operation. These are:

PART	V SUPPLY	DESCRIPTION
AD7224	+ 12 to 15V	CMOS 8 bit, voltage out
AD7225	+ 12 to 15V	CMOS quad 8 bit, voltage out
AD7226	+ 12 to 15V	CMOS quad 8 bit, voltage out

1 MICROAMP DAC



The operating current of most CMOS DACs is very dependent on the logic level of the digital input signals. This is because internal input level shifters account for most of the DAC's power consumption. If the digital inputs are made to swing from V_{DD} to ground (i.e. CMOS logic) rather than over smaller TTL levels, the supply current can typically be cut by a factor of 100 or more in the steady state. This does not reduce power consumption when switching, since the supply current transiently rises to 5 to 10mA for 10 to 20ns, independent of input logic level.

MAXIM'S CMOS DAC IMPROVEMENTS

AD75XX Series DACs – Reduced Parasitic Capacitance

- C_{OUT} Reduced from 200pF to 100pF
- Lower DAC Glitch Energy

AD7225/7226 Voltage Output – Sources AND Sinks Current

- Drives Capacitive Loads
- Improved Output Settling Time – 1.5us (typical)
- Equal Settling with Single or Dual Supplies

Voltage References

PART NUMBER	OUTPUT VOLTAGE	TEMPERATURE COEFFICIENT (ppm/°C)	INITIAL OUTPUT VOLTAGE ACCURACY	SUPPLY VOLTAGE	SUPPLY CURRENT (TYP/MAX mA)	FEATURES
MAX670	10V	3	±2.5mV	+15V	9/14	Kelvin Sensing
MAX671	10V	1	±1.0mV	+15V	9/14	Kelvin Sensing
AD580	2.5V	10	±10mV	+15V	0.75/1.0	+5V Power
AD581	10V	5	±5mV	+15V	0.75/1.0	
AD584	10V, 7.5V, 5V, 2.5V	5	±5mV	+15V	0.75/1.0	Programmable Output Voltage
AD2700	10V	3	±2.5mV	+15V	9/14	Precision
AD2710	10V	1	±1.0mV	+15V	9/14	Precision
AD2701	-10V	2	±2.5mV	-15V	9/14	Precision
ICL8069	1.23V	10 to 100	±25mV	-	50µA to 5mA	2 Terminal Bandgap Reference
REF01E	10V	8.5	±30mV	+15V	1/1.4	
REF01HP	10V	25	±50mV	+15V	1/1.4	
REF02E	5V	8.5	±15mV	+15V	1/1.4	
REF02HP	5V	25	±25mV	+15V	1/1.4	

Integrating A/D Converter Selector

PART NUMBER	RESOLUTION	OUTPUT TYPE	SUPPLY VOLTAGE	SUPPLY CURRENT (TYP/MAX mA)	TYPICAL APPLICATIONS AND COMMENTS
MAX133	3 ³ / ₄ Digit ±4000 Counts	µP	9V	0.09/0.2	Digital Multimeters
MAX134	3 ³ / ₄ Digit ±4000 Counts	µP	+5V	0.09/0.2	Weigh Scales, Instruments
MAX136	3 ¹ / ₂ Digit ±2000 Counts	LCD	9V	0.06/0.15	Hold Function, Low Power
ICL7106	3 ¹ / ₂ Digit ±2000 Counts	LCD Drive	9V	0.6/1.8	Digital Multimeters
ICL7107	3 ¹ / ₂ Digit ±2000 Counts	LED Drive	±5V	0.6/1.8	Digital Panel Meters
ICL7109	12 Bits + Sign ±4096 Counts	8/16 bit µP and UART	±5V	0.7/1.5	Data loggers, Process Control Up to 30 conversions/second
ICL7116	3 ¹ / ₂ Digit ±2000 Counts	LCD	9V	0.8/1.8	Same as ICL7106, but adds Hold function
ICL7117	3 ¹ / ₂ Digit ±2000 Counts	LED	±5V	0.8/1.8	Same as ICL7107, but adds Hold function
ICL7126	3 ¹ / ₂ Digit ±2000 Counts	LCD	9V	0.09/0.2	Use ICL7136 for new designs
ICL7129 ICL7129A	4 ¹ / ₂ Digit ±20,000 Counts	Triplexed LCD	9V	1.0/1.4	DPMs, Instruments Lowest Noise A/D - 3µV (7129A)
ICL7135	4 ¹ / ₂ Digit ±20,000 Counts	Multiplexed BCD	±5V	1.0/2.0	DMM, DPM, Data Loggers
ICL7136	3 ¹ / ₂ Digit ±2000 Counts	LCD	9V	0.06/0.1	Low Power version of ICL7106 Very Low Noise
ICL7137	3 ¹ / ₂ Digit ±2000 Counts	LED	±5V	0.06/0.2	Low Power when LED display turned off

INTEGRATING vs SAR A/D

Parameter	Integrating A/D (MAX134)	SAR (AD7572)
Conversion Time	50ms	5 μ s
Conversions/Sec	20	200,000
Resolution	16+ Bits	12 Bits
Normal Mode Rejection of 50/60Hz	50+ dB	0dB
Power Consumption	1mW	135mW
Typical Applications	Data Loggers, Digital Multimeters, High Resolution, Slow Measurements	High Speed Instruments, Digital Signal Processing

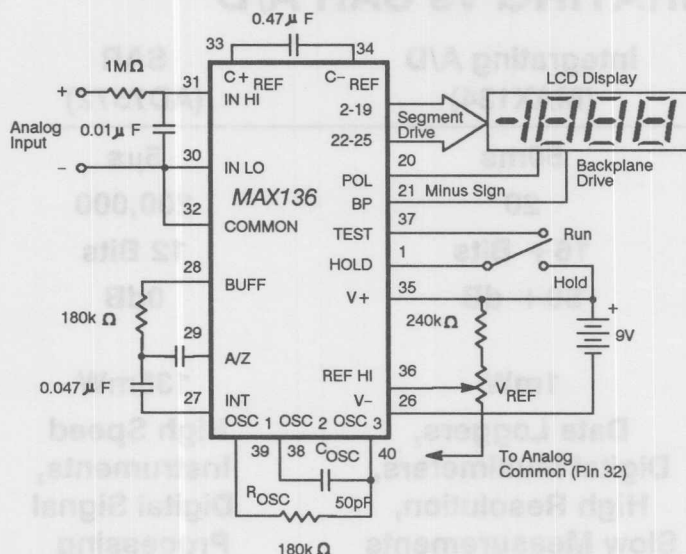
MAX133 AND MAX134 A/Ds FOR INSTRUMENTS

- $\pm 40,000$ Count Resolution
- 20 Conversions Per Second
- Microprocessor Interface
- 100 μ A Supply Current
- 7 Input Channels
- Onboard Switches for 5 Decade Attenuator, Ohms, Current, and AC Ranges
- 10 μ V Resolution

The MAX133 and MAX134 combine both low power operation and high resolution with a versatile input routing circuit that can be configured as either a 5 decade attenuator or as a 7 channel multiplexer. Since it is an integrating A/D converter, it has excellent rejection of power line frequencies, and in addition has the op amp and switching network to add an active 2 pole filter for additional AC rejection.

The MAX133 and MAX134 do not have display drive capability, instead they have a bus interface suitable for connection to either 4 bit or 8 bit microprocessors. The MAX133 and MAX134 differ slightly in that the MAX133 has a multiplexed address/data bus to reduce to a minimum the number of I/O pins used by the microprocessor to read and control the A/D. The MAX134 has a separate 4 bit bidirectional data bus and 3 address inputs, simplifying the hardware and software interface with 8 bit microprocessors.

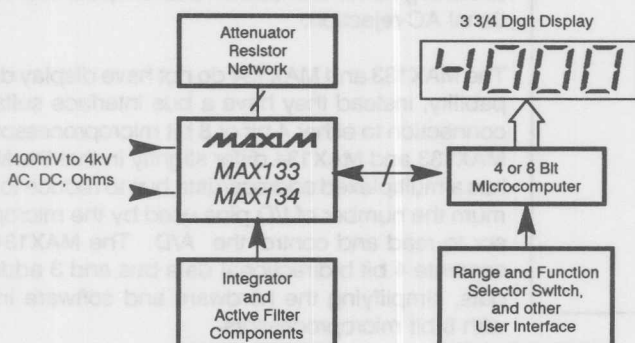
"SINGLE CHIP" INSTRUMENT



This is the complete circuit of a typical LCD digital panel meter. All DPM manufacturers who use this IC provide essentially the same features, since the basic functional operation of the end product is

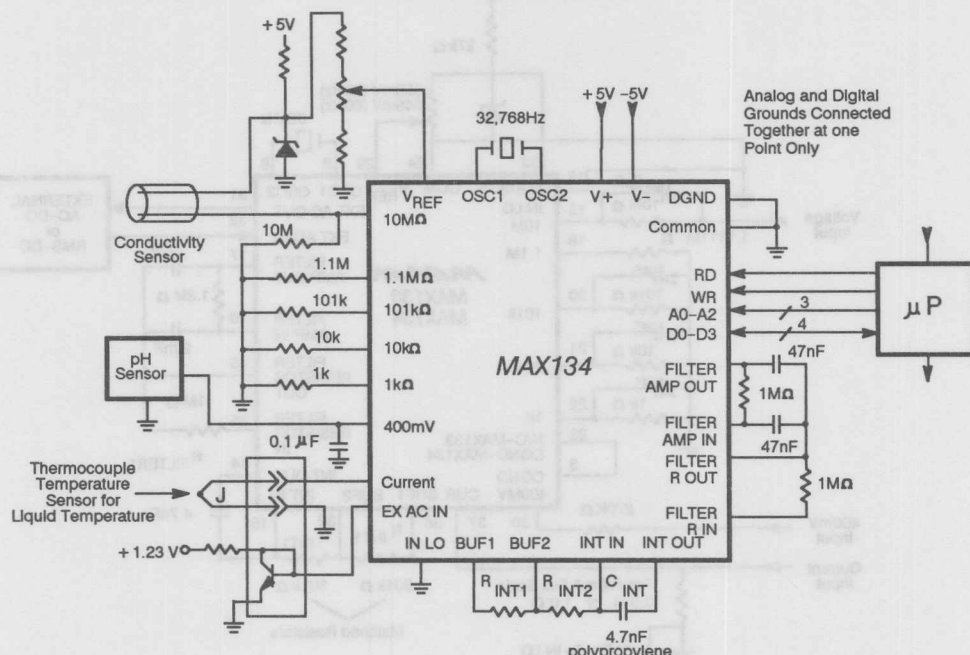
predefined by the IC manufacturer. Enhanced features such as auto-calibration are very difficult to add.

"2 CHIP" INSTRUMENT



This block diagram shows how the MAX133 and MAX134 are designed to work in conjunction with a microprocessor/display driver. The measurement function and the processing/display of data have been partitioned into two different ICs. Through the use of different software, the same set of hardware can provide radically different features. Since these features are defined primarily by the software, the instrument manufacturer rather than the A/D manufacturer determines the functional operation of the instrument. It is relatively simple with this system to add enhanced features such as auto-calibration, scaling of the display in different units, alarm limits, and minimum/maximum peak recording.

ANALOG FRONT END FOR pH, CONDUCTIVITY, AND TEMPERATURE MEASUREMENT



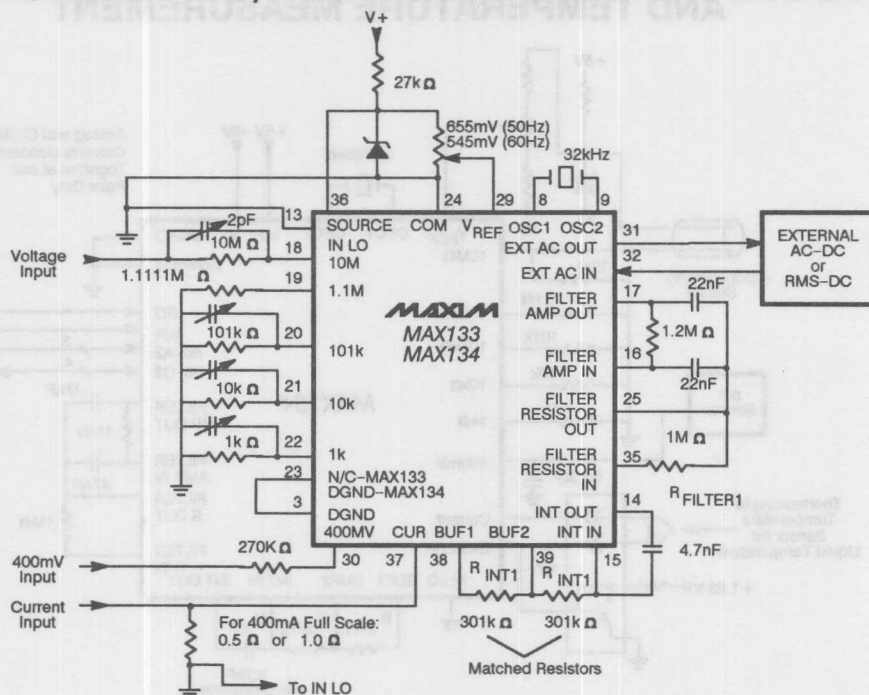
This combination pH/Conductivity/Temperature meter provides a complete chemical profile of a unknown liquid. The temperature sensor is a type J thermocouple which has an output of about 53μV/°C. The thermocouple output voltage depends on the "cold junction" temperature, which in this system is the temperature of the isothermal thermocouple connection block on the front panel of the meter. The temperature of this block is measured via a simple diode temperature sensor. The microprocessor uses a multisegment lookup table and interpolation to convert the A/D output code to the corresponding temperature display.

The pH sensor output is zero for a neutral solution of 7pH, and changes about -59mV per pH unit at 25°C. The precise voltage change per pH unit varies from -54mV/pH unit when the pH sensor temperature is 0°C, to -74mV/pH unit with a sensor temperature of 100°C. The microprocessor easily compensates for this 30% change in gain, since it already measures the actual temperature of the liquid. A small lookup table then gives the gain correction by which the measured pH is multiplied to obtain the temperature cor-

rected pH. This temperature correction is a good example of the advantage of a 2 chip instrument over an instrument based on a single-chip A/D.

Without changing hardware, the instrument manufacturer also can add the ability to exactly calibrate the pH probe. A typical method is to have the user place the pH probe into two or three buffer solutions of known pH, whose pH values are entered into the instrument via a keyboard. The microprocessor then generates offset and gain calibration corrections and stores these sensor corrections in battery backed up CMOS RAM or EEPROM. Maximum limits for the correction factors are normally set to detect gross errors in the sensor or, more often, errors in the keyboard entry of the pH values of the buffer solutions. If the standard buffer solutions are always the same pH (typically 4, 7, and 14pH), the user does not even need to enter the buffer pH, but need only put the system into the calibration mode. Once the calibration mode is entered, the meter will recognize which of the three buffer solutions is present and will automatically generate a new set of calibration constants.

BASIC MAX133/MAX134 ANALOG CIRCUIT

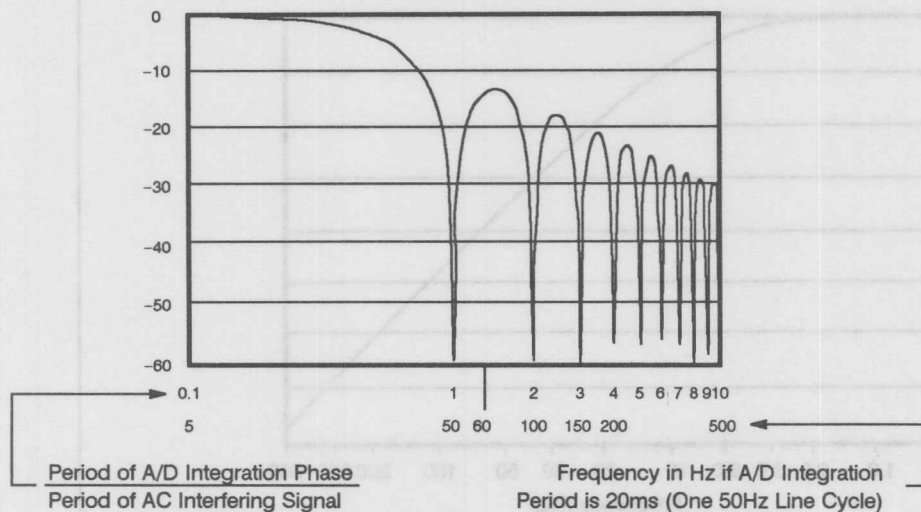


This is the analog portion of a digital multimeter. The precision attenuator resistors on the left side are connected to a set of kelvin-sensed analog switches which are controlled by the microprocessor through bits in the MAX133 control registers. The analog switches are controlled by the microprocessor to select 400mV, 4V, 40V, 400V, and 4000V ranges. The external AC-DC converter can be bypassed for DC measurements or inserted into the signal path

for AC measurements, again under the control of the microprocessor. Similarly, the selection of the current input and control of the active filter are also controlled by the microprocessor.

The MAX133/134 does all of the critical timing related to the actual A/D conversion, but higher level control and setup functions are controlled by the microprocessor via the MAX133 control register bits.

(SIN X)/X REJECTION



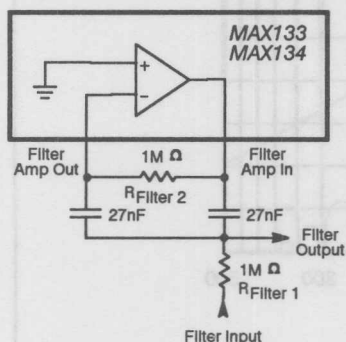
Rejection of 50/60Hz line frequencies is very important since the signals being measured are very small and inadvertent pickup of 50/60Hz signals often means that there are several counts of 50/60Hz interference riding on top of the signal, both as common mode interference and differential or normal mode interference. The integrating A/D will have essentially zero response to these 50/60Hz signals provided the integration period is an integral multiple of the interfering signal. In particular, if the integration period is chosen to be exactly one period of the line frequency, the line frequency and all of its harmonics will be rejected since the integral of a sine wave over one or more complete cycles is zero, regardless of the starting phase. Simultaneous rejection of both 50Hz and 60Hz can be achieved with an integration time of 100ms, which is 6 complete cycles of 60Hz and 5 complete cycles of 50Hz. With standard dual slope A/D converters such as the ICL7107 and ICL7135 fami-

lies, this results in a total conversion time of 400ms, or 2.5 conversions per second.

The MAX133 and MAX134 provide rejection of either 50Hz or 60Hz (but not both simultaneously) with a constant clock frequency by adjusting the number of clock cycles in the integration period in response to a register bit controlled by the microprocessor. This allows the MAX133/134 to achieve 50/60Hz rejection while still providing 20 conversions per second.

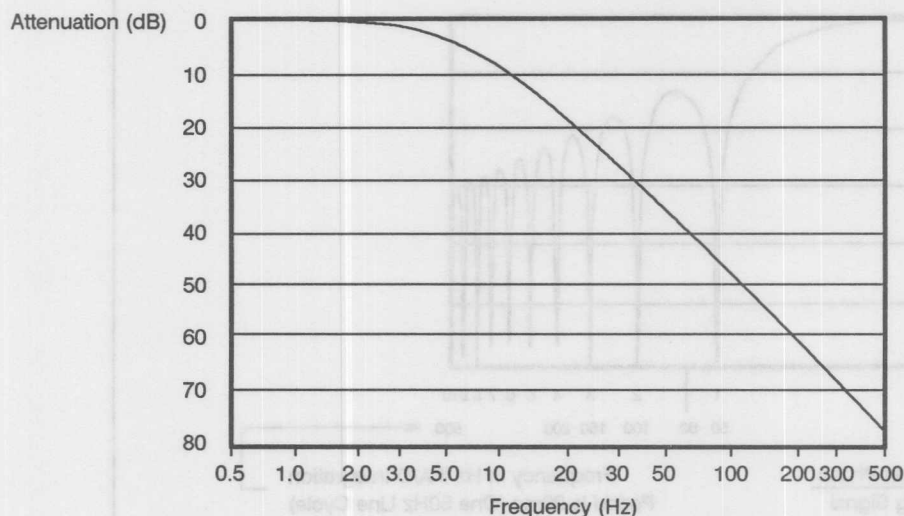
Rejection of AC signals which are not synchronous with the power line is achieved in all integrating A/Ds through the inherent low pass characteristic of the integrator. The combination of synchronous and asynchronous rejection of AC signals follows the $(\sin x)/x$ characteristic.

MAX133/134 2 POLE ACTIVE FILTER WITH ZERO OFFSET

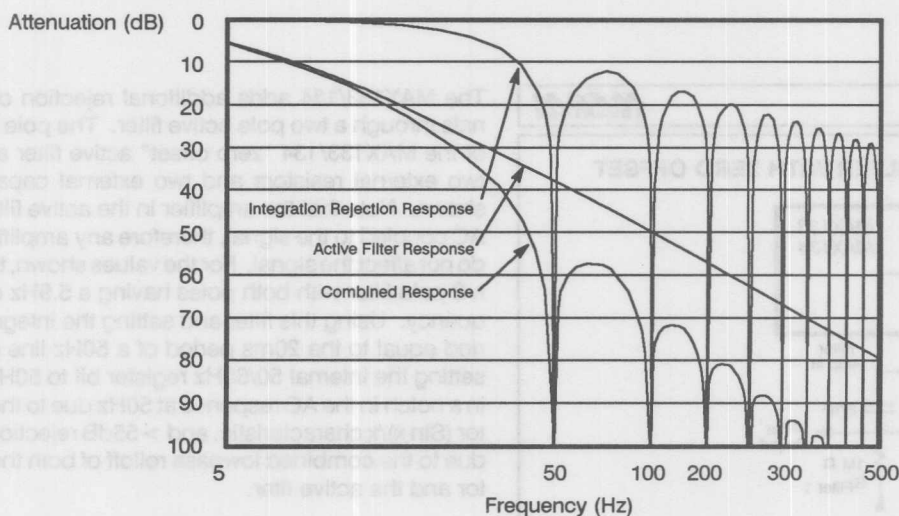


The MAX133/134 adds additional rejection of AC signals through a two pole active filter. The pole locations of the MAX133/134 "zero offset" active filter are set by two external resistors and two external capacitors as shown. Note that the amplifier in the active filter is only AC coupled to the signal, therefore any amplifier offsets do not affect the signal. For the values shown, the filter is a 2 pole filter with both poles having a 5.9Hz cutoff frequency. Using this filter and setting the integration period equal to the 20ms period of a 50Hz line cycle (by setting the internal 50/60Hz register bit to 50Hz) results in a notch in the AC response at 50Hz due to the integrator $(\sin x)/x$ characteristic, and > 55dB rejection at 60Hz due to the combined lowpass rolloff of both the integrator and the active filter.

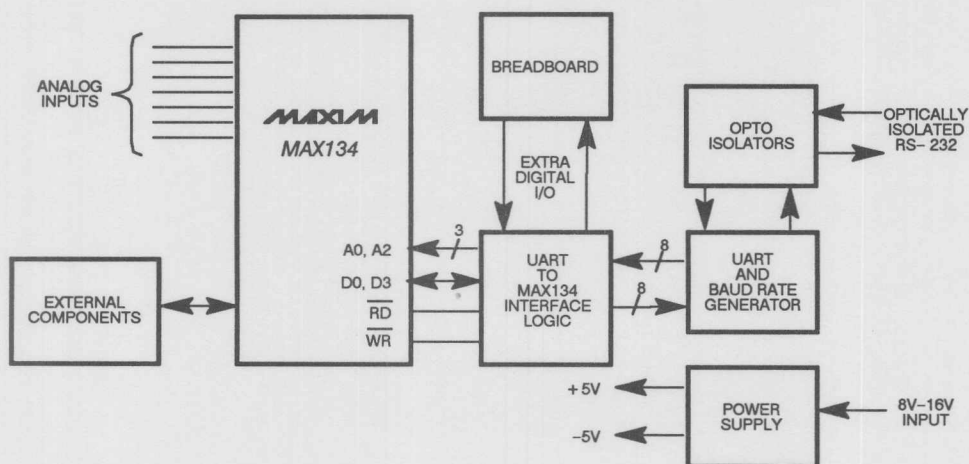
RESPONSE OF MAX133/134 TWO POLE ACTIVE FILTER



COMBINED FREQUENCY RESPONSE



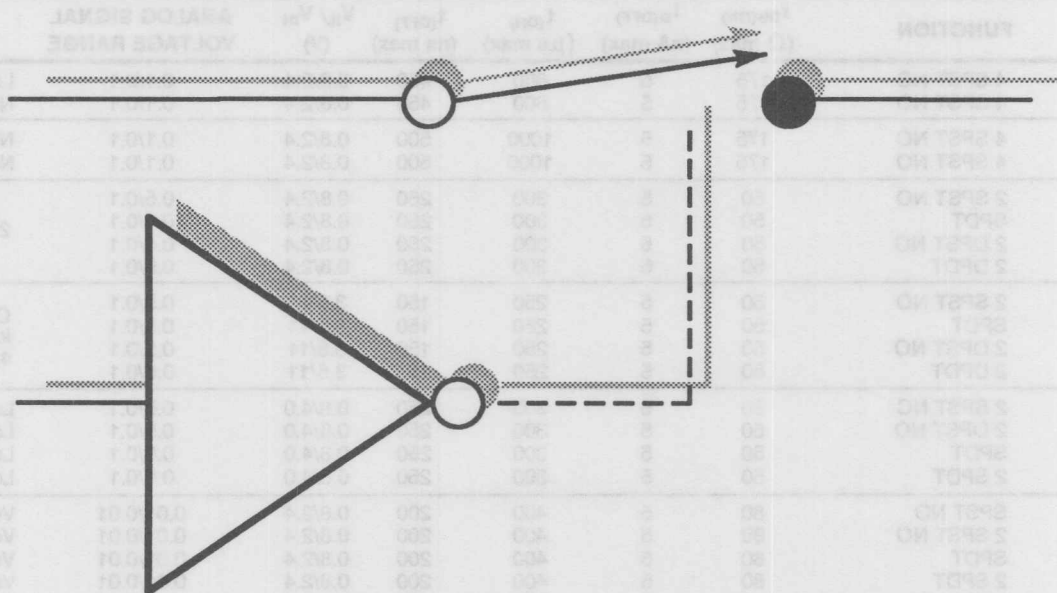
MAX134 EVALUATION KIT BLOCK DIAGRAM



A/D performance is generally characterized by speed, conversion noise, integral linearity, and differential linearity. Normally these are specified by the IC manufacturer, though not necessarily for the exact set of operating conditions of each possible application. Secondary characteristics such as overload recovery and time to stabilize after power is turned on are often not specified at all. This means that in most cases real, live hardware must be used to check the suitability of a given A/D for a particular application. Maxim has

simplified this evaluation process by making available a demonstration board which converts the MAX134 digital output to an optoisolated RS-232 serial data link. A/D evaluation can be as simple as plugging the RS-232 link into an IBM PC or compatible computer since the evaluation kit comes complete with software drivers. A breadboarding area is including for building input conditioning circuitry unique to a particular application.

23-2116-1000A



Analog Multiplexers

PART NUMBER	FUNCTION	$r_{DS(ON)}$ (Ω max)	$I_{D(OFF)}$ (nA max)	$t_{(ON)}$ (μ s max)	V_{IL}/V_{IH} (V)	ANALOG SIGNAL VOLTAGE RANGE	FEATURES
DG508A DG509A	1 of 8 2 of 8	300	2	1 μ s	0.8/2.4	± 15 V	Industry standard
HI508A HI509A	1 of 8 2 of 8	1500	2	1 μ s	0.8/2.4	-12.5V to +13.5V	Fault Protected
IH5108A IH6108 IH6208	See MAX358 See DG508A See DG509A						
MAX310 MAX311	1 of 8 2 of 8	250	10	1.5 μ s	0.8/2.4	± 15 V	70dB isolation at 10MHz
MAX312 MAX313	1 of 8 2 of 8	250	10	1.5 μ s	0.8/2.4	± 15 V	Video mux with latches
MAX358 MAX359	1 of 8 2 of 8	1500	2	1 μ s	0.8/2.4	-12.5V to +13.5V	Fault Protected to
MAX368 MAX369	1 of 8 2 of 8	1500	2	1 μ s	0.8/2.4	-12.5V to +13.5V	Fault Protected, with address latches

Analog Switches

PART NUMBER	FUNCTION	$r_{DS(ON)}$ (Ω max)	$I_{D(OFF)}$ (nA max)	$t_{(ON)}$ (μ s max)	$t_{(OFF)}$ (ns max)	V_{IL}/V_{IH} (V)	ANALOG SIGNAL VOLTAGE RANGE	FEATURES
DG201A DG202	4 SPST NC 4 SPST NO	175 175	5 5	600 600	450 450	0.8/2.4 0.8/2.4	0.1/0.1 0.1/0.1	Low power Normally open
DG211 DG212	4 SPST NC 4 SPST NO	175 175	5 5	1000 1000	500 500	0.8/2.4 0.8/2.4	0.1/0.1 0.1/0.1	No V_{LOGIC} Supply Normally open
DG300A DG301A DG302A DG303A	2 SPST NO SPDT 2 DPST NO 2 DPDT	50 50 50 50	5 5 5 5	300 300 300 300	250 250 250 250	0.8/2.4 0.8/2.4 0.8/2.4 0.8/2.4	0.5/0.1 0.5/0.1 0.5/0.1 0.5/0.1	2.4V _{IH} , Low R_{ON}
DG304A DG305A DG306A DG307A	2 SPST NO SPDT 2 DPST NO 2 DPDT	50 50 50 50	5 5 5 5	250 250 250 250	150 150 150 150	3.5/11 3.5/11 3.5/11 3.5/11	0.5/0.1 0.5/0.1 0.5/0.1 0.5/0.1	CMOS Logic levels, high speed, Low R_{ON}
DG381A DG384A DG387A DG390A	2 SPST NC 2 DPST NO SPDT 2 SPDT	50 50 50 50	5 5 5 5	300 300 300 300	250 250 250 250	0.8/4.0 0.8/4.0 0.8/4.0 0.8/4.0	0.5/0.1 0.5/0.1 0.5/0.1 0.5/0.1	Low R_{ON} Low R_{ON} Low R_{ON} Low R_{ON}
IH5040 IH5041 IH5042 IH5043 IH5044 IH5045	SPST NO 2 SPST NO SPDT 2 SPDT DPST NO 2 DPST NO	80 80 80 80 80 80	5 5 5 5 5 5	400 400 400 400 400 400	200 200 200 200 200 200	0.8/2.4 0.8/2.4 0.8/2.4 0.8/2.4 0.8/2.4 0.8/2.4	0.01/0.01 0.01/0.01 0.01/0.01 0.01/0.01 0.01/0.01 0.01/0.01	Very low power Very low power Very low power Very low power Very low power Very low power
IH5048 IH5049 IH5050 IH5051	SPDT NO 2 SPST NO SPDT 2 SPDT	45 45 45 45	5 5 5 5	1000 1000 1000 1000	500 500 500 500	0.8/2.4 0.8/2.4 0.8/2.4 0.8/2.4	0.01/0.01 0.01/0.01 0.01/0.01 0.01/0.01	Low charge injection
IH5140 IH5141 IH5142 IH5143 IH5144 IH5145	SPST NO 2 SPST NO SPDT 2 SPDT DPST NO 2 DPST NO	50 50 50 50 50 50	0.1 0.1 0.1 0.1 0.1 0.1	150 150 200 200 200 200	125 125 125 125 125 125	0.8/2.4 0.8/2.4 0.8/2.4 0.8/2.4 0.8/2.4 0.8/2.4	0.01/0.01 0.01/0.01 0.01/0.01 0.01/0.01 0.01/0.01 0.01/0.01	Fast; low power Fast; low power Fast; low power Fast; low power Fast; low power Fast; low power
IH5341 IH5352	2 SPST NO 4 SPST NO	75 75	0.5 0.5	300 300	150 150	0.8/2.4 0.8/2.4	0.001/0.001 0.001/0.001	70dB isolation at 10MHz
MAX331 MAX332	4 SPST NC 4 SPST NO	175 175	5 5	600 600	450 450	0.8/2.4 0.8/2.4	0.01/0.01 0.01/0.01	Improved DG201A/DG202

Analog Switches and Multiplexers

High Voltage Switches

- Flying Capacitor Level Translator

Fault Protected Analog Multiplexers

- How They Work
- Redundant Systems

Maxim's Analog Switch Families

Family	Typical Devices	Features
General Purpose CMOS	DG201, IH5040	$\pm 15V$ Operation
High Voltage CMOS	MAX340, MAX343	$\pm 50V$ Operation
Fault Protected Mux	MAX358, MAX359	$\pm 35V$ Fault Protected
Video Switches	IH5341, IH5352	70dB Isolation at 10MHz
Video Multiplexer	MAX310, MAX311	75dB Isolation at 5MHz, 1 of 8 and 2 of 8 mux

MAX341-348 High Voltage Switches

- Power Supplies: $\pm 50V$ or $+60V/Gnd$
- On Resistance: 80Ω with MAX348
- Analog Signal Range: $V+ \text{ to } V-$
- Supply Current: $300\mu A$
- Switching time: $< 1\mu s$

MAX343

V_{IN}

$\pm 50V$ Common Mode

$+50V$

$-50V$

$1\mu F$

$50pF^*$

$400Hz$
 $0-15V$

$1\mu F$

V_{OUT}

$50pF^*$

* Optional Charge Injection Compensation Capacitor

This flying capacitor level translator can remove up to $\pm 50\text{V}$ common mode voltage from small differential signals, with up to 120dB of common mode rejection ratio. First this circuit charges the “flying capacitor” (the 1uF capacitor on the left side of the circuit) with the differential voltage. Next, the two sets of switches connect the flying capacitor to the 1uF output capacitor. The output capacitor can be referenced to any voltage from -50V to $+50\text{V}$. In this case, the output is referenced to ground. The differential voltage across the flying capacitor charges the output capacitor to the same differential voltage that appears between the two differential input terminals at the left side of the circuit.

This circuit is normally used to remove large common mode signals from DC or slow varying signals from such sensors such as temperature sensors. The MAX343 analog switch does allow operation of the flying capacitor circuit at a switching rates greater than 100kHz, making it suitable for signals up to 1kHz if desired.

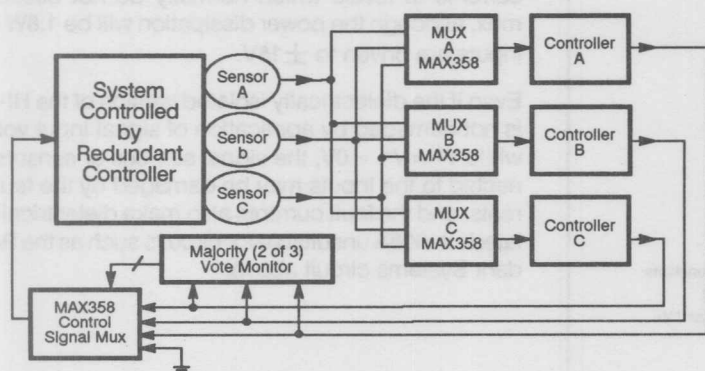
This circuit is a very cost effective alternative to isolation amplifiers if an operating common mode voltage range of $\pm 50\text{V}$ is acceptable. Simply adding a current limiting resistor in series with each of the input leads will protect the circuit against common mode voltage of several hundred volts, although the output voltage is correct only if both of the inputs are within the $\pm 50\text{V}$ range.

FLYING CAPACITOR LEVEL TRANSLATOR

Common Mode Range	±50V
CMRR	110dB to 120dB (depends on switching frequency and capacitor value)
Bandwidth	Low, B.W. < < Sampling Frequency
Power Dissipation	25mW

The two main errors are caused by leakage currents and charge injection. Charge injection from the gates to the sources and drains of the analog switches pumps a small amount of charge into the output each time the switches change state. The optional 50pF capacitors compensate for this charge injection, increasing CMRR to greater than 120dB, or 1 μ V of output voltage error for each volt of common mode voltage change.

REDUNDANT CONTROL SYSTEM



In systems such as fly-by-wire avionics control systems, and control systems for large chemical plants it is desirable to have both redundant sensors and redundant control logic. When using conventional multiplexers, each of the redundant controllers must have its own complete set of sensors, since failure of the power supply in the controller causes a conventional mux to short the sensor to ground.

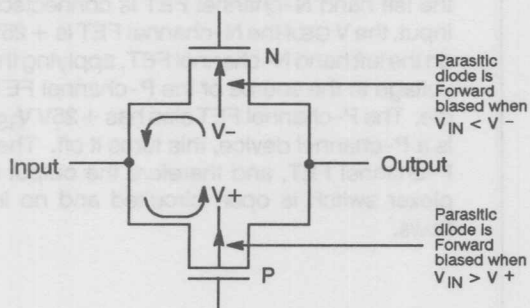
The MAX358 fault protected multiplexer maintains a high input impedance whether or not it is powered, and therefore allows one set of sensors to provide inputs to all redundant controllers. Alternatively, the MAX358 allows redundant sensors to be connected to all controllers without fear of total system disruption through the failure of the power supply of just one of the controllers.

The outputs of the controllers are monitored by a majority vote circuit, which compares the outputs of the controllers and drives the controlled system with either one of the three controller outputs or a failsafe signal. The MAX358 mux at the bottom left corner is used for this control signal selection since any overvoltage output of up to $\pm 35\text{V}$ from any one of the controllers will not affect the output when another controller is selected. Even if a controller with an overvoltage output of up to $\pm 35\text{V}$ is selected, the output of the MAX358 will remain within $\pm 13.5\text{V}$.

MAX358 Fault Protected Multiplexer

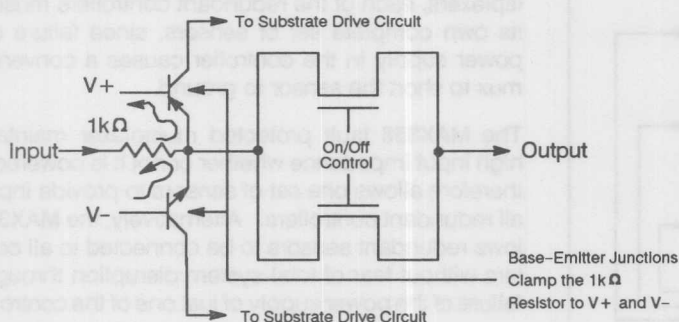
- Enhanced Versions of HI-508A, HI-509A
- 1 of 8 Channels -- MAX358
- 2 of 8 Channel Mux -- MAX359
- $\pm 35\text{V}$ (70Vp-p) Fault Protection
- Only Nanoamperes of Input Current
- --- under all fault conditions
- All Switches are Off when Power Supply is Off
- Operates from $\pm 4.5\text{V}$ to $\pm 18\text{V}$ Supplies

FAULT CURRENTS IN STANDARD CMOS DG508



The parasitic diodes between the body and the source/drain of a CMOS FET provide low impedance paths to V^+ and V^- . While Maxim uses proprietary current limiting circuitry to prevent device destruction when $V^+ = V^- = 0\text{V}$ and the inputs are powered, a fault current of 20mA will flow. Other manufacturers devices do not have current limiting circuitry and may damage both the signal source and the analog switch itself if voltages greater than a couple of volts are applied when $V^+ = V^- = 0\text{V}$.

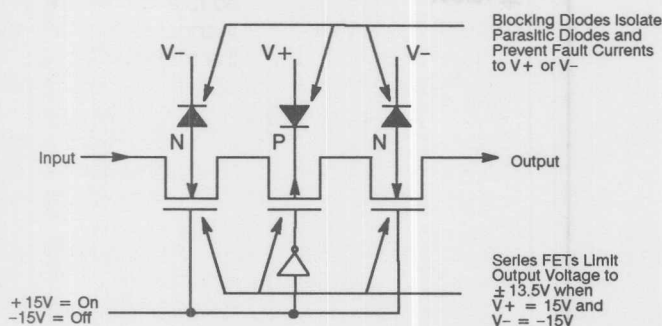
FAULT CURRENTS IN D.I. HI508A



The "fault protected" dielectrically isolated version of the HI-508A limits fault currents by putting a nominal $1k\Omega$ resistor in series with the inputs. This limits the fault currents to levels which normally do not destroy the mux, although the power dissipation will be $1.8W$ if all 8 inputs are driven to $\pm 15V$.

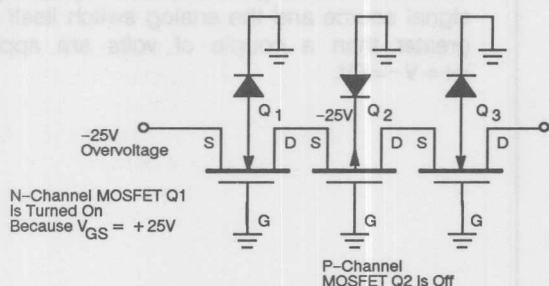
Even if the dielectrically isolated version of the HI-508A is not damaged by application of signal input voltages while $V^+ = V^- = 0V$, the signal sources or sensors connected to the inputs may be damaged by the fault currents, and the fault currents also make dielectrically isolated HI-508A unsuitable for circuits such as the Redundant Systems circuit above.

FAULT PROTECTED MUX USING JUNCTION ISOLATED CMOS, MAX358



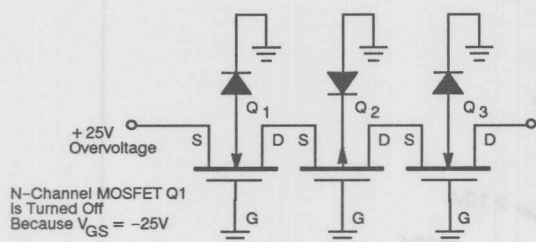
The MAX358 has a blocking diode in series with each of the parasitic body-to-drain/source diodes. This limits the current through the parasitic diodes to just picoamps or nanoamps of leakage. Unlike standard analog switches which have a P-channel and an N-channel FET in parallel, the MAX358 has 2 N-channel FETs and a P-channel FET in series. As shown in the following four drawings, this prevents input fault currents, even if the output is grounded and $V^+ = V^- = 0V$. The three FET series structure also limits the output voltage swing to $\pm 13.5V$ when the MAX358 is powered by $\pm 15V$ supplies, even if the selected input channel swings $\pm 35V$.

-25V OVERVOLTAGE WITH MULTIPLEXER POWER OFF



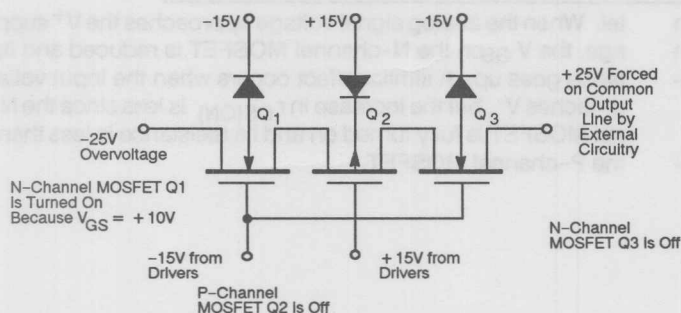
This is the schematic of one of the eight switches in the MAX358 8 channel multiplexer. When $V^+ = V^- = 0V$, the gates of all three FETs are at ground. Since the source of the left hand N-channel FET is connected to the -25V input, the V_{GS} of the N-channel FET is $+25V$. This turns on the left hand N-channel FET, applying the -25V input voltage to the source of the P-channel FET in the middle. The P-channel FET also has $+25V V_{GS}$, but since it is a P-channel device, this turns it off. The drain of the P-channel FET, and therefore the output of the multiplexer switch is open circuited and no input current flows.

+ 25V OVERVOLTAGE WITH MULTIPLEXER POWER OFF



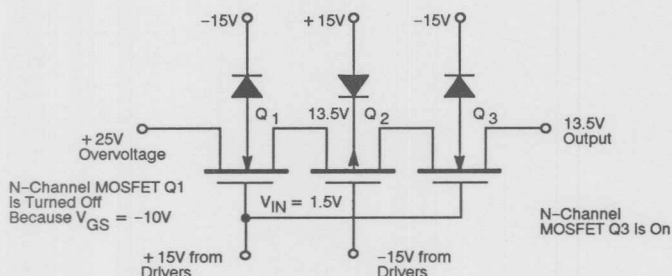
When a + 25V overvoltage is applied while $V^+ = V^- = 0V$, the V_{GS} of the left hand N-channel FET is -25V and it turns off. Only a few picoamps leakage current flows into the input.

-25V OVERVOLTAGE ON AN OFF CHANNEL WITH MULTIPLEXER POWER ON



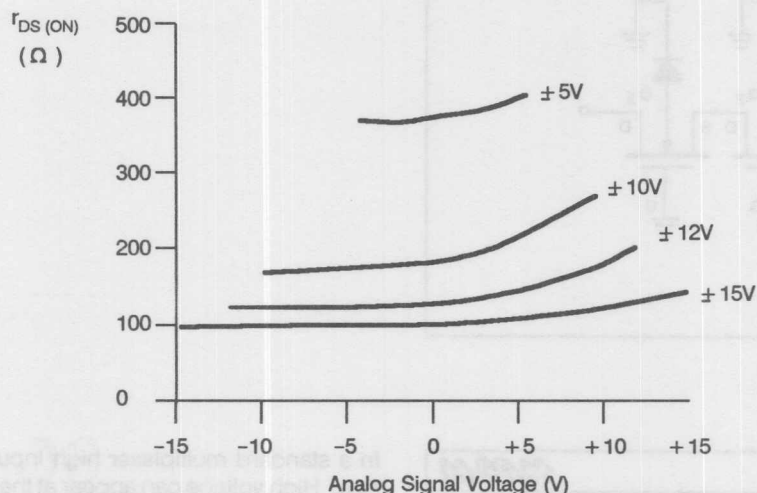
In a standard multiplexer high input fault currents will flow. High voltage can appear at the multiplexer output, possibly damaging sensitive circuitry such as S/D converters which are connected to the output of the mux. The MAX358 limits or clips the output voltage to $\pm 13.5V$. With -25V applied to a channel which is not selected, the left hand N-channel device is turned on, but the P-channel FET in the middle has a +40V V_{GS} and is turned off, isolating the overvoltage input. If a positive overvoltage is applied, either to the input or to the output, the N-channel FET becomes a source follower, and voltage at the other terminal of the N-channel FET will be clipped to an N-channel threshold below V^+ . Since the N-channel threshold is 1.5 to 4V, the output will be clipped to between 13.5V and 11V when $V^+ = 15V$. This voltage is applied to the P-channel FET. Since the P-channel FET in this case is turned off by a +15V gate drive, the +13.5V signal will not pass through the P-channel FET. In other words, not only will the MAX358 not be damaged by a +25V overvoltage on an unselected channel, that overvoltage will not interfere with normal operation.

+ 25V OVERVOLTAGE INPUT TO THE ON CHANNEL



When an overvoltage is applied to the input channel selected by the address decoder of the multiplexer, the series FET structure limits the output voltage to +13.5V. When a positive overvoltage is applied to the selected channel, the N-channel device becomes a source follower limiting the positive output voltage to $(V^+ - N\text{-channel Threshold})$. Negative overvoltages are clipped to a P-channel threshold more positive than V^- . This clipping or voltage limiting action is bilateral, and overvoltages applied to the common output of the multiplexer are clipped to $\pm 13.5V$ when the MAX358 is used as a 1 channel to 8 channel demultiplexer.

DG201 ON RESISTANCE

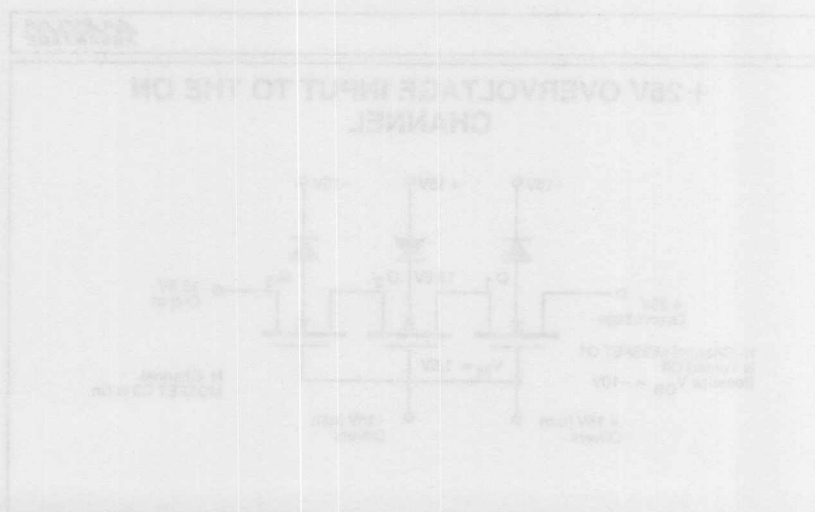


The on resistance or $r_{DS(ON)}$ of an analog switch varies with both the power supply voltages and the analog input voltage. In both cases, the $r_{DS(ON)}$ varies because the V_{GS} (gate-to-source voltage) changes.

The DG201 uses an N-channel and a P-channel MOSFET in paral-

lel. When the analog signal voltage approaches the V^+ supply voltage, the V_{GS} on the N-channel MOSFET is reduced and its resistance goes up. A similar effect occurs when the input voltage approaches V^- , but the increase in $r_{DS(ON)}$ is less since the N-channel MOSFET is fully turned on and its resistance is less than that of the P-channel MOSFET.

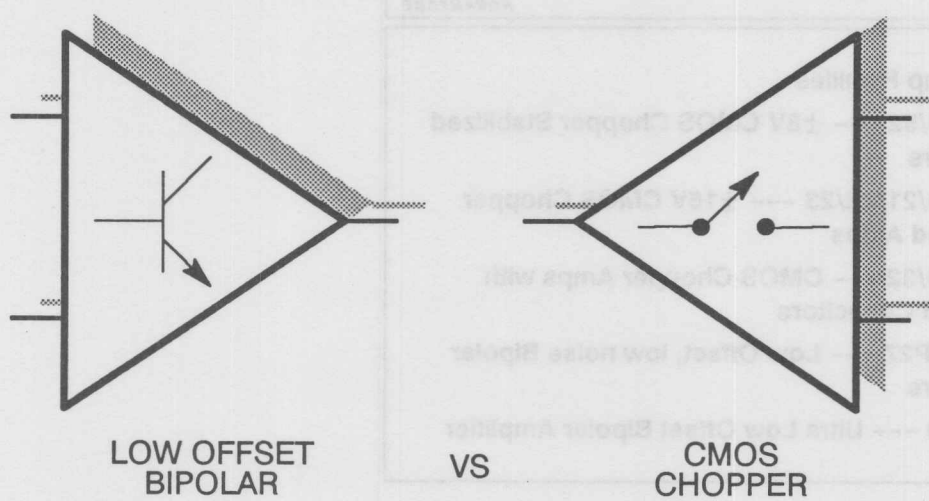
When an overvoltage is applied to the input channel, the output voltage is limited to the supply voltage. The P-channel MOSFET is turned on and its resistance is low. When a positive overvoltage is applied to the output, the N-channel MOSFET is turned on and its resistance is low. When a negative overvoltage is applied to the output, the P-channel MOSFET is turned on and its resistance is low. The output voltage is limited to the supply voltage. The P-channel MOSFET is turned on and its resistance is low. When a positive overvoltage is applied to the output, the N-channel MOSFET is turned on and its resistance is low. When a negative overvoltage is applied to the output, the P-channel MOSFET is turned on and its resistance is low. The output voltage is limited to the supply voltage.



PART NUMBER	INITIAL VOS (V max)	TEMP-OS (V/°C max)	PSRR (dB min)	SUPPLY VOLTAGE (V max)	SUPPLY CURRENT (mA max)	NOISE DC-1kHz (V rms/√Hz)	FEATURES
MAX4220	2	0.05	10	±5.5V to ±15V	2.0	0.3	Low Iq
MAX4220B	10	0.1	20	±5.5V to ±15V	2.0	0.7	Lowest Iq
MAX4220C	2	0.02	30	±5.5V to ±15V	2.0	0.5	Lowest Iq
MAX4220D	10	0.05	30	±5.5V to ±15V	2.0	0.5	+15V Operation
MAX4220E	10	0.05	30	±5.5V to ±15V	2.0	0.5	+15V Operation
MAX4221	10	0.05	30	±5.5V to ±15V	2.0	0.5	Low Iq
MAX4221B	10	0.05	30	±5.5V to ±15V	2.0	0.5	Low Iq
MAX4221C	10	0.05	30	±5.5V to ±15V	2.0	0.5	No 1st Order
MAX4221D	10	0.05	30	±5.5V to ±15V	2.0	0.5	No 2nd Order
MAX4221E	10	0.05	30	±5.5V to ±15V	2.0	0.5	No 3rd Order
MAX4221F	10	0.05	30	±5.5V to ±15V	2.0	0.5	No 4th Order
MAX4221G	10	0.05	30	±5.5V to ±15V	2.0	0.5	No 5th Order
MAX4221H	10	0.05	30	±5.5V to ±15V	2.0	0.5	No 6th Order
MAX4221I	10	0.05	30	±5.5V to ±15V	2.0	0.5	No 7th Order
MAX4221J	10	0.05	30	±5.5V to ±15V	2.0	0.5	No 8th Order
MAX4221K	10	0.05	30	±5.5V to ±15V	2.0	0.5	No 9th Order
MAX4221L	10	0.05	30	±5.5V to ±15V	2.0	0.5	No 10th Order
MAX4221M	10	0.05	30	±5.5V to ±15V	2.0	0.5	No 11th Order
MAX4221N	10	0.05	30	±5.5V to ±15V	2.0	0.5	No 12th Order
MAX4221O	10	0.05	30	±5.5V to ±15V	2.0	0.5	No 13th Order
MAX4221P	10	0.05	30	±5.5V to ±15V	2.0	0.5	No 14th Order
MAX4221Q	10	0.05	30	±5.5V to ±15V	2.0	0.5	No 15th Order
MAX4221R	10	0.05	30	±5.5V to ±15V	2.0	0.5	No 16th Order
MAX4221S	10	0.05	30	±5.5V to ±15V	2.0	0.5	No 17th Order
MAX4221T	10	0.05	30	±5.5V to ±15V	2.0	0.5	No 18th Order
MAX4221U	10	0.05	30	±5.5V to ±15V	2.0	0.5	No 19th Order
MAX4221V	10	0.05	30	±5.5V to ±15V	2.0	0.5	No 20th Order
MAX4221W	10	0.05	30	±5.5V to ±15V	2.0	0.5	No 21st Order
MAX4221X	10	0.05	30	±5.5V to ±15V	2.0	0.5	No 22nd Order
MAX4221Y	10	0.05	30	±5.5V to ±15V	2.0	0.5	No 23rd Order
MAX4221Z	10	0.05	30	±5.5V to ±15V	2.0	0.5	No 24th Order

MAXIM
ADVANTAGE

PRECISION OP-AMP COMPARISON



Precision Operational Amplifiers

PART NUMBER	INITIAL VOS (μV max)	VOS TEMPCO ($\mu\text{V}/^\circ\text{C}$ max)	I_{BIAS} (pA max)	SUPPLY VOLTAGE	SUPPLY CURRENT (mA max)	NOISE DC-1Hz (μV pk-pk typ)	FEATURES
ICL7650	5	0.05	10	$\pm 2.25\text{V}$ to $\pm 8\text{V}$	2.0	0.7	Low I_{BIAS}
ICL7650B	10	0.1	20	$\pm 2.25\text{V}$ to $\pm 8\text{V}$	2.0	0.7	Lowest Cost
ICL7652	5	0.05	30	$\pm 2.5\text{V}$ to $\pm 8\text{V}$	2.0	0.2	Lowest Noise
MAX420	10	0.05	30	$\pm 2.5\text{V}$ to $\pm 16.5\text{V}$	2.0	0.3	$\pm 15\text{V}$ Operation
MAX421	10	0.05	30	$\pm 2.5\text{V}$ to $\pm 16.5\text{V}$	2.0	0.3	
MAX422	10	0.05	30	$\pm 2.5\text{V}$ to $\pm 16.5\text{V}$	0.5	0.4	Low Current
MAX423	10	0.05	30	$\pm 2.5\text{V}$ to $\pm 16.5\text{V}$	0.5	0.4	
MAX430	10	0.05	30	$\pm 2.5\text{V}$ to $\pm 16.5\text{V}$	2.0	0.3	No Ext Capacitors
MAX432	10	0.05	30	$\pm 2.5\text{V}$ to $\pm 16.5\text{V}$	0.5	0.4	No Ext Capacitors
MAX400C	15	0.3	2000	$\pm 3\text{V}$ to $\pm 18\text{V}$	4.0	0.15	Non-Chopped
OP07A	25	0.6	2000	$\pm 3\text{V}$ to $\pm 18\text{V}$	4.0	0.15	Non-Chopped
OP07E	75	1.3	4000	$\pm 3\text{V}$ to $\pm 18\text{V}$	4.0	0.15	Non-Chopped
OP27A	25	0.6	40,000	$\pm 4\text{V}$ to $\pm 18\text{V}$	4.7	0.18	Non-Chopped
OP27E	25	0.6	40,000	$\pm 4\text{V}$ to $\pm 18\text{V}$	4.7	0.18	Non-Chopped



Maxim Op Amp Families

- ICL7650/52 --- $\pm 8\text{V}$ CMOS Chopper Stabilized Amplifiers
- MAX420/21/22/23 --- $\pm 15\text{V}$ CMOS Chopper Stabilized Amps
- MAX430/32 --- CMOS Chopper Amps with Onboard Capacitors
- OP07/OP27 --- Low Offset, low noise Bipolar Amplifiers
- MAX400 --- Ultra Low Offset Bipolar Amplifier

PRECISION OP-AMP COMPARISON

	MAX430 (CMOS Chopper)	MAX400 (Low V_{OS} Bipolar)
Offset	5 μ V	15 μ V
Drift	0.05 μ V/ $^{\circ}$ C	0.3 μ V/ $^{\circ}$ C
10 Hz Noise	1.1 μ Vp-p	0.35 μ Vp-p
0.1Hz Noise	0.1 μ V	0.2 μ V
Input I Bias	0.1nA	2nA
Speed	0.5MHz, 0.5V/ μ s	0.6MHz, 0.3V/ μ s
Clock Ripple	20 μ Vp-p (400Hz)	none

Both CMOS chopper stabilized amplifiers and bipolar op-amps have benefited from constant improvement over time. As an advance in the performance of each type is made, it is immediately declared "superior" to the other. In reality, both have their advantages, and we attempt to make these clearer here.

The table outlines some significant spec differences between state-of-the-art examples of each type. This is a true "apples-to-apples" comparison in that the CMOS chopper amp, the MAX430, operates from ± 15 V power and requires no external zeroing capacitors. It is therefore drop-in compatible with the MAX400, a premium grade precision bipolar op-amp.

As can be seen, the MAX430 holds an advantage with regard to offset voltage and drift while the MAX400 has superior 10Hz noise and of course no clock ripple since it is unchopped. Not mentioned in the table but sometimes significant is the MAX400's superior recovery time from input overloads compared to that of chopper stabilized op-amps.

THE MAX430 IS NOT A TSC911

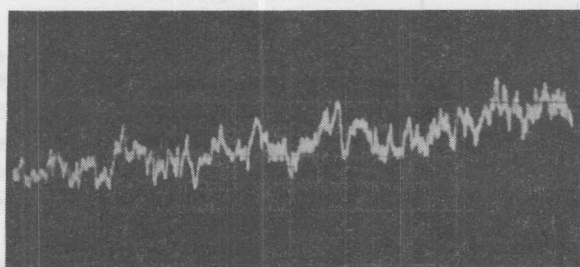
- Much Larger Internal Capacitors
- Much Lower Noise

10Hz Measured Noise

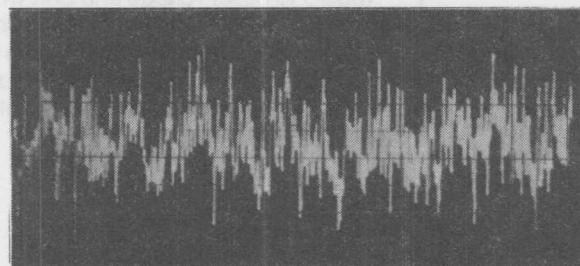
MAX430	TSC911
1.1 μ Vp-p	8 μ Vp-p

The MAX430 is NOT similar to a TSC911. Although both are CMOS chopper stabilized op-amps, and both do not need external auto-zero capacitors, the similarity ends there. The internal capacitors in the MAX430 are over 700 times larger than those used in the TSC911. This gives the MAX430 eight times better noise performance. In most applications that call for chopper stabilized amplifiers, noise is the next most important parameter after offset voltage and drift. The MAX430 also will operate from up to ± 15 V power while the TSC911 is limited to ± 7 V.

1 Hz NOISE



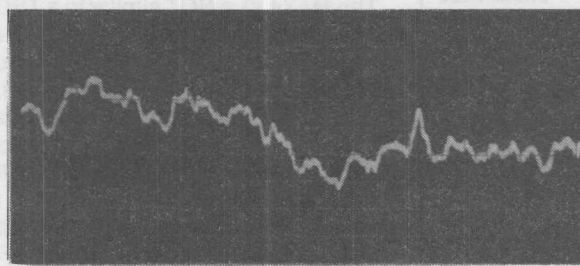
OP07/MAX400



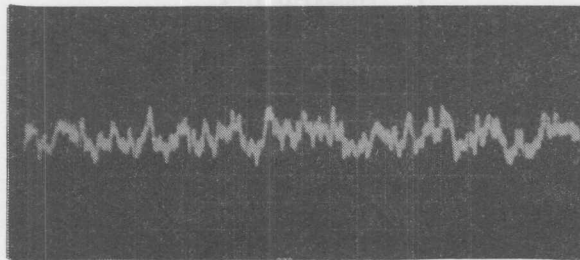
MAX420

Vert: 0.1 μ V/div
Horiz: 5s/div

0.1 Hz NOISE



OP07/MAX400



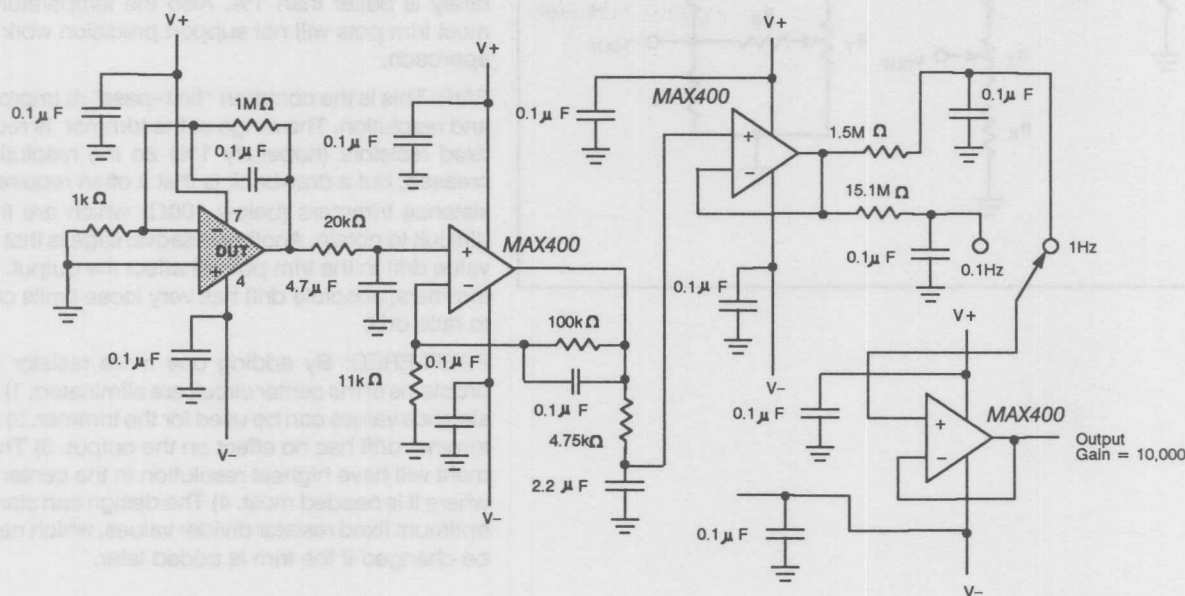
MAX420

Vert: 0.1 μ V/div
Horiz: 20s/div

The MAX430 and MAX420's main advantage over precision bipolar op-amps comes to light when working with very low bandwidth, low level signals, such as those from strain gauges and load cells. Though the bipolar amp is generally regarded as the lower noise part, and is so in a 1Hz or greater bandwidth, the CMOS chopper provides a lower usable noise floor below 1Hz, where 1/f noise becomes significant. The bipolar amp's noise actually rises while the chopped amplifier effectively cancels 1/f.

These four photos compare the noise performance of a MAX400 (bipolar) to a MAX420 (CMOS chopper) in a 1Hz and 0.1Hz bandwidth. The MAX400 clearly shows lower peak-to-peak noise in the 1Hz photo, although some baseline drift, about 130nV, is visible. The length of the horizontal axis in the 1Hz photo is 50 seconds. In the 0.1Hz photo the MAX420 shows lower peak-to-peak noise over a 3 minute period.

NOISE TEST CIRCUIT



The circuit used for the preceding photographs has an overall gain of 10,000. A gain of 1000, taken at the device under test (DUT), is followed by a gain of 10. The schematic differs from some common noise test circuits in that the measurement bandwidth used here extends to DC.

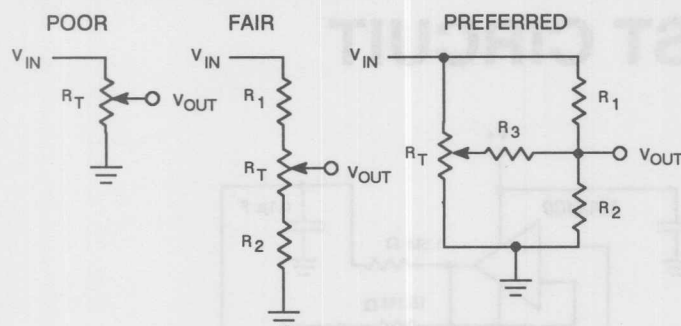
Noise specs are typically quoted from 0.1Hz to the upper limit frequency to stay away from offset drift with temperature and 1/f noise.

Offset drift and 1/f noise are lumped together in these photos because most real applications of precision op-amps are not afforded the luxury of separating the sources of low frequency error. Performance down to DC is what usually matters. Both bipolar and CMOS chopper amplifiers do very well when their strengths are utilized.

ACHIEVING LOW DRIFT PERFORMANCE

Device	Input Offset Voltage (mV)	Input Bias Current (pA)	Input Impedance (M Ω)
MAX400	0.1	10	> 100
MAX422	0.1	10	> 100
MAX423	0.1	10	> 100
MAX424	0.1	10	> 100
MAX425	0.1	10	> 100
MAX426	0.1	10	> 100
MAX427	0.1	10	> 100
MAX428	0.1	10	> 100
MAX429	0.1	10	> 100
MAX430	0.1	10	> 100
MAX431	0.1	10	> 100
MAX432	0.1	10	> 100
MAX433	0.1	10	> 100
MAX434	0.1	10	> 100
MAX435	0.1	10	> 100
MAX436	0.1	10	> 100
MAX437	0.1	10	> 100
MAX438	0.1	10	> 100
MAX439	0.1	10	> 100
MAX440	0.1	10	> 100
MAX441	0.1	10	> 100
MAX442	0.1	10	> 100
MAX443	0.1	10	> 100
MAX444	0.1	10	> 100
MAX445	0.1	10	> 100
MAX446	0.1	10	> 100
MAX447	0.1	10	> 100
MAX448	0.1	10	> 100
MAX449	0.1	10	> 100
MAX450	0.1	10	> 100
MAX451	0.1	10	> 100
MAX452	0.1	10	> 100
MAX453	0.1	10	> 100
MAX454	0.1	10	> 100
MAX455	0.1	10	> 100
MAX456	0.1	10	> 100
MAX457	0.1	10	> 100
MAX458	0.1	10	> 100
MAX459	0.1	10	> 100
MAX460	0.1	10	> 100
MAX461	0.1	10	> 100
MAX462	0.1	10	> 100
MAX463	0.1	10	> 100
MAX464	0.1	10	> 100
MAX465	0.1	10	> 100
MAX466	0.1	10	> 100
MAX467	0.1	10	> 100
MAX468	0.1	10	> 100
MAX469	0.1	10	> 100
MAX470	0.1	10	> 100
MAX471	0.1	10	> 100
MAX472	0.1	10	> 100
MAX473	0.1	10	> 100
MAX474	0.1	10	> 100
MAX475	0.1	10	> 100
MAX476	0.1	10	> 100
MAX477	0.1	10	> 100
MAX478	0.1	10	> 100
MAX479	0.1	10	> 100
MAX480	0.1	10	> 100
MAX481	0.1	10	> 100
MAX482	0.1	10	> 100
MAX483	0.1	10	> 100
MAX484	0.1	10	> 100
MAX485	0.1	10	> 100
MAX486	0.1	10	> 100
MAX487	0.1	10	> 100
MAX488	0.1	10	> 100
MAX489	0.1	10	> 100
MAX490	0.1	10	> 100
MAX491	0.1	10	> 100
MAX492	0.1	10	> 100
MAX493	0.1	10	> 100
MAX494	0.1	10	> 100
MAX495	0.1	10	> 100
MAX496	0.1	10	> 100
MAX497	0.1	10	> 100
MAX498	0.1	10	> 100
MAX499	0.1	10	> 100

GENERAL PURPOSE TRIM TECHNIQUES



This may seem like an elementary subject, but were betting 1/2 page that it isn't. The tighter the tolerances required in a given analog design, the more likely that one or more trims will be needed. From left to right are three progressively improved trim circuits:

POOR: The cheapest, but suffers from poor resolution. Even with high quality multi-turn trimmers, settability rarely is better than 1%. Also the temperature drift of most trim pots will not support precision work with this approach.

FAIR: This is the common "first-pass" at improving drift and resolution. The range of the trimmer is reduced by fixed resistors (hopefully 1%) so the resolution is increased, but a drawback is that it often requires low resistance trimmers (below 100 Ω) which are frequently difficult to obtain. Another disadvantage is that absolute value drift in the trim pot will affect the output. On most trimmers, absolute drift has very loose limits compared to ratio drift.

PREFERRED: By adding one more resistor (R_3), the problems of the center circuit are eliminated: 1) Large resistance values can be used for the trimmer. 2) Absolute trimmer drift has no effect on the output. 3) The adjustment will have highest resolution in the center of range where it is needed most. 4) The design can start with the optimum fixed resistor divider values, which needed not be changed if the trim is added later.

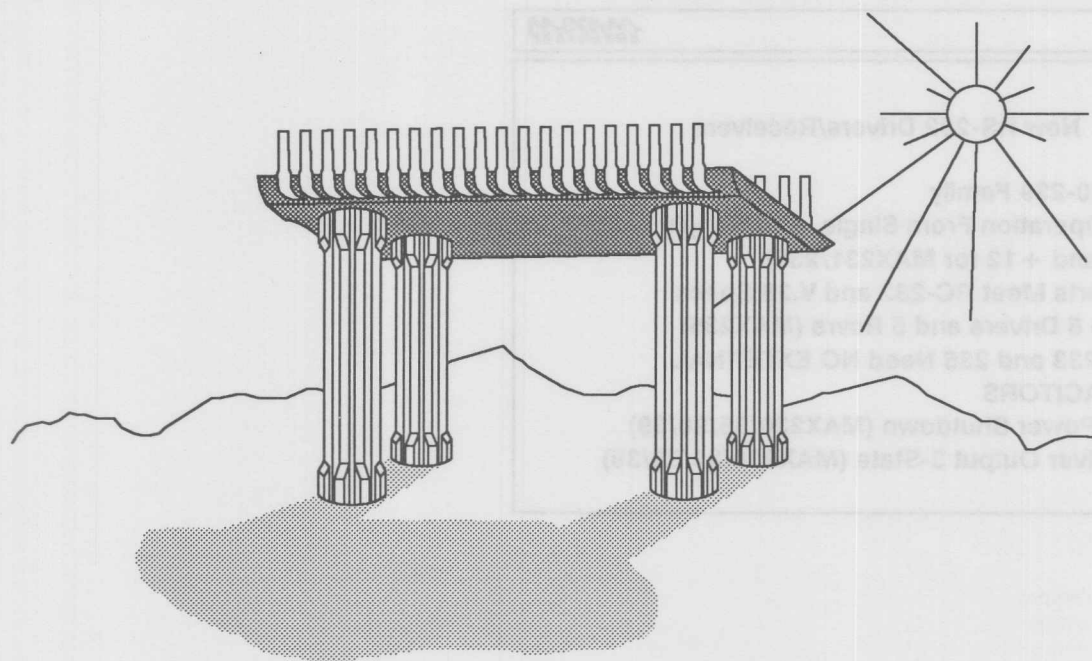
ACHIEVING LOW DRIFT PERFORMANCE

Type	R Tempco Absolute	(ppm/ $^{\circ}$ C) Tracking	Manufacturers
Metal Film	25 to 100	—	Mepco, General, Resistance, Corning, TRW
Wire Wound	1 to 20	—	Kelvin, Julie Research, Ultronix
Special Films and Foils	< 1 to 20	to 0.5	Vishay, Ultronix, Caddock
Networks	5 to 25	0.5 to 5	Allen Bradley, Caddock, Vishay

Picking a good low drift, low noise op-amp is not all there is to precision analog design, although it's not a bad start.

A critical concern in most precision applications, along with amplifier offset drift and noise, is gain error and drift. The key to minimizing gain drift usually rests with resistor selection. The table lists common precision resistor types and their temperature coefficients.

MICROPROCESSOR SUPPORT CIRCUITS



PART NUMBER	POWER SUPPLY VOLTAGE	NO. OF DRIVERS	NO. OF RECEIVERS	EXTERNAL COMPONENTS	LOW POWER SHUTDOWN AND/OR TTL STATE	NO. OF PINS
MAX300	+5V	8	8	4 Capacitors	Yes	50
MAX301	+5V and +1.5V to 1.8V	8	8	2 Capacitors	No	48
MAX302	+5V	8	8	4 Capacitors	No	48
MAX303	+5V	8	8	None	No	50
MAX304	+5V	8	8	1 Capacitor	No	48
MAX305	+5V	8	8	None	Yes	52
MAX306	+5V	8	8	4 Capacitors	Yes	52
MAX307	+5V	8	8	4 Capacitors	No	52
MAX308	+5V	8	8	4 Capacitors	No	52
MAX309	+5V and +1.5V to 1.8V	8	8	2 Capacitors	Yes	52

Microprocessor Support Products/Applications

- New RS-232 Products
- RS-232 Applications
- MAX680 +5V to $\pm 10V$ Converter
- MAX690/691 Microprocessor Supervisor

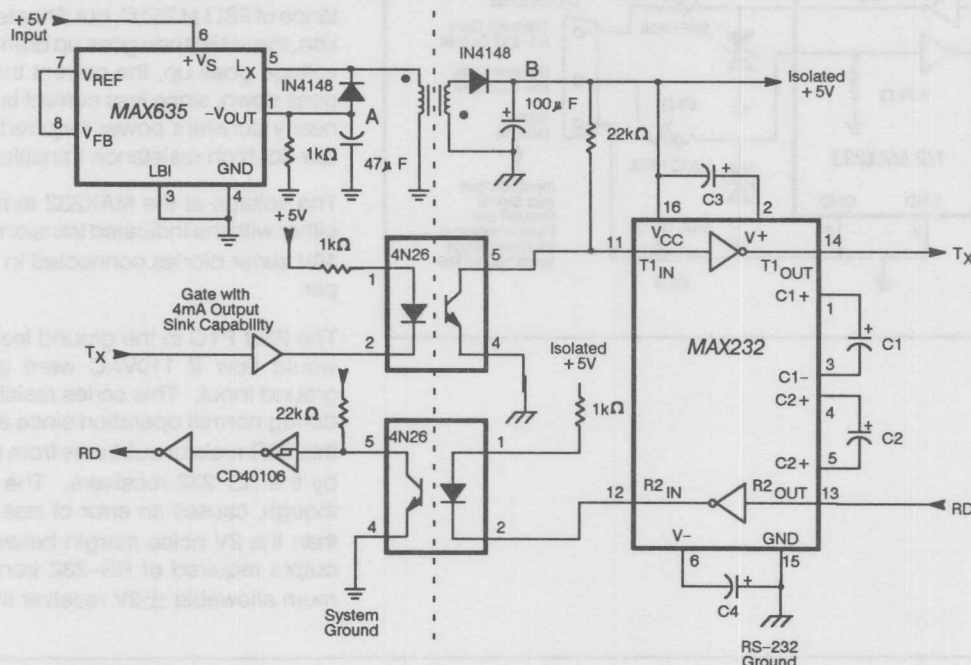
New RS-232 Drivers/Receivers
The MAX230-239 Family

- Full Operation From Single +5V Supply (+5 and +12 for MAX231/239)
- All Parts Meet RC-232 and V.28 Specs
- Up to 5 Drivers and 5 Rcvrs (MAX235)
- MAX233 and 235 Need NO EXTERNAL CAPACITORS
- Low Power Shutdown (MAX230/35/36/39)
- Receiver Output 3-State (MAX230/35/36/39)

RS-232 Drivers and Receivers

PART NUMBER	POWER SUPPLY VOLTAGE	NO. OF RS-232 DRIVERS	NO. OF RS-232 RECEIVERS	EXTERNAL COMPONENTS	LOW POWER SHUTDOWN AND/OR TTL 3-STATE	NO. OF PINS
MAX230	+5V	5	0	4 Capacitors	Yes	20
MAX231	+5V and +7.5V TO 13.2V	2	2	2 Capacitors	No	14
MAX232	+5V	2	2	4 Capacitors	No	16
MAX233	+5V	2	2	None	No	20
MAX234	+5V	4	0	4 Capacitors	No	16
MAX235	+5V	5	5	None	Yes	24
MAX236	+5V	4	3	4 Capacitors	Yes	24
MAX237	+5V	5	3	4 Capacitors	No	24
MAX238	+5V	4	4	4 Capacitors	No	24
MAX239	+5V and +7.5V to 13.2V	3	5	2 Capacitors	Yes	24

OPTO-ISOLATED RS-232 INTERFACE

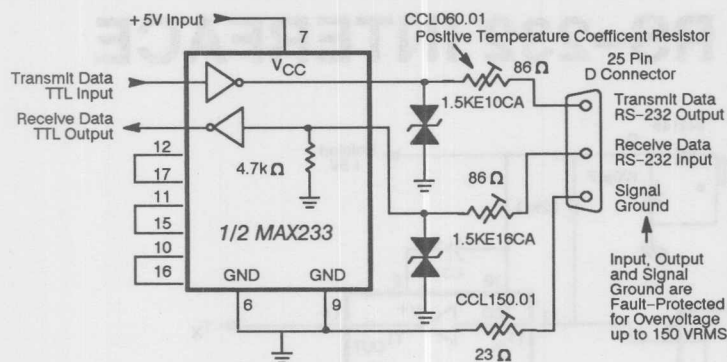


Although the RS-232 specification recommends that the maximum cable length should be less than 15 meters, much longer cable lengths are fairly common. A problem sometimes encountered, particularly when using long cables, is that ground potential differences between the two ends of the cable can be higher than the 2V noise margin built into the RS-232 specification. Another possibility that must sometimes be considered is the potentially disastrous effects of accidental connection of the RS-232 inputs or outputs to high voltages such as the 110/220VAC power line. Optoisolation both isolates the computer or instrument from possibly harmful volt-

ages and allows operation with a large voltage difference between the grounds at each end of an RS-232 link.

The MAX635 DC-DC converter IC provides an isolated power supply by regulating the output of the transformer primary. With a bifilar wound (1 to 1 turns ratio) transformer, when the output of the primary is regulated to 5V, the output of the secondary will be semi-regulated to 5V. Line and load regulation is easily kept to $\pm 10\%$ if the primary and secondary are tightly coupled by using bifilar winding techniques. The data is transferred via the two 4N26 optocouplers.

FAULT PROTECTED RS-232 INTERFACE

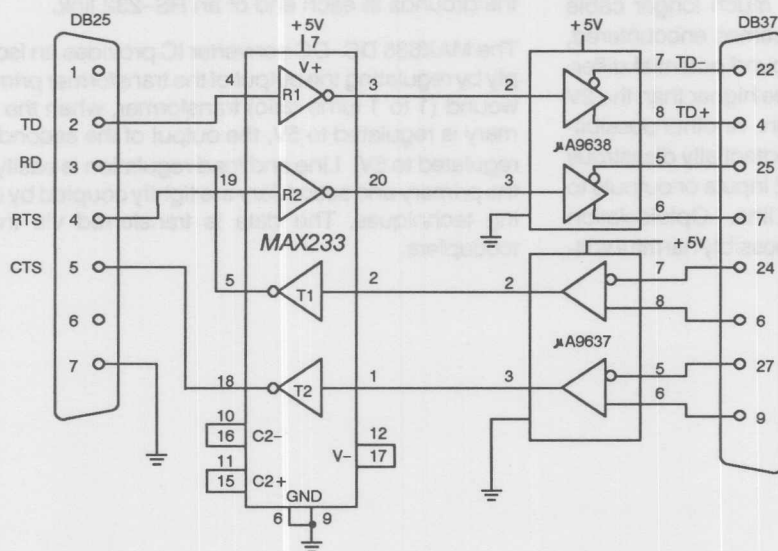


By combining positive temperature coefficient current limiting resistors with the back-to-back series diodes or transient suppressors, the RS-232 inputs and outputs of the MAX232 can be protected against connection to 110VAC. The CCL060.01 PTC from Midwest Components (Muskegon, Michigan, USA) has a nominal resistance of 86Ω at 25°C , but if heated by I^2R power dissipation, the resistance goes up dramatically. As the applied voltage goes up, the current through the PTC actually goes down, since less current is needed to maintain the nearly constant power required to keep the PTC at its low-to-high resistance transition temperature.

The voltage at the MAX232 terminals can be clamped either with the indicated transient suppressors, or by two 10V zener diodes connected in series as a $\pm 11\text{V}$ clipper.

The 23Ω PTC in the ground lead limits the current that would flow if 110VAC were applied to the RS-232 ground input. This series resistance must be kept low during normal operation since any voltage drop across this 23Ω resistor subtracts from the RS-232 signal seen by the RS-232 receivers. The 23Ω series resistance, though, causes an error of less than 50mV, much less than the 2V noise margin between the minimum $\pm 5\text{V}$ output required of RS-232 transmitters and the maximum allowable $\pm 3\text{V}$ receiver thresholds.

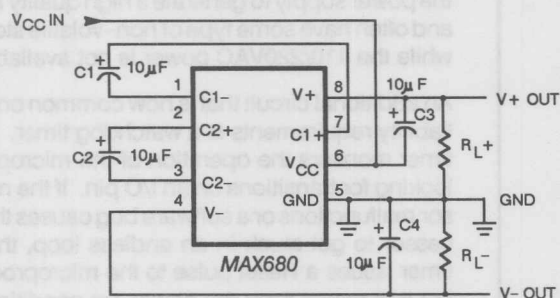
RS-232 TO RS-422 CONVERTER



Although most terminals, computers, and printers use the RS-232 (CCITT V.24) standard for their serial communication port, some equipment uses the RS-422 or RS-485 balanced or differential standard to achieve a higher data rate, or to accommodate up to $\pm 5\text{V}$ voltage difference between the ground potentials between the two pieces of equipment which communicate via RS-422. This cir-

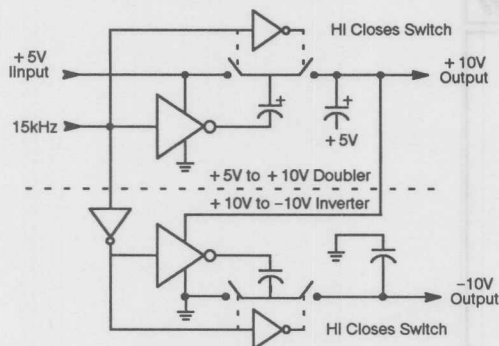
cuit, which uses only a single +5V supply, accepts RS-232 inputs and converts them to RS-422 outputs. In the other direction, it accepts RS-422 inputs and converts them to RS-232 outputs. The MAX233 converts the +5V power supply input to $\pm 9\text{V}$, using 4 internal capacitors.

+5V TO $\pm 10V$ VOLTAGE CONVERTER



The MAX23X series of RS-232 transmitter/receiver ICs drive the transmitter outputs to RS-232/V.24 levels while requiring only a single +5V supply. They are able to do this because they have a DC-DC converter circuit which doubles the +5V input to +10V and then inverts that +10V supply to -10V. The MAX680 is an 8 pin IC which has just the +5V to $\pm 10V$ DC-DC converter.

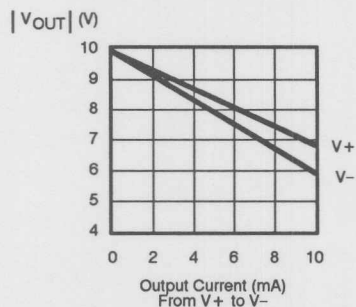
MAX680 AND MAX23X CHARGE PUMP CIRCUIT



Both the MAX680 and MAX23X use a capacitive charge pump technique for the voltage conversion. When the 15kHz internal clock is high, the output of the inverter in the +5V to +10V doubler pulls the capacitor to ground, while the left hand switch connects the other terminal of the capacitor to +5V. When the +15kHz clock goes low, the inverter output drives the negative terminal of the capacitor to +5V. The positive terminal of the capacitor is still +5V higher than the negative terminal, which means that it is at +10V. The +10V on the capacitor positive terminal is routed to the +10V output filter capacitor via the right hand switch.

The +10V to -10V charge pump voltage inverter works similarly, except that the CMOS inverter swings from ground to +10V rather than from ground to +5V.

OUTPUT VOLTAGE vs. OUTPUT CURRENT FROM V+ TO V-



Since the charge pump voltage inverter is not regulated, the full $\pm 10V$ output is available only at very low output currents. The output impedance of the V+ output is about 300 Ω , as shown by the 3V drop to a +7V output when 10mA is drawn from the V+ terminal. The V- output has about a 400 Ω output impedance. The quiescent current of the MAX680 is 1mA.

MICROPROCESSOR SUPPORT FUNCTIONS

Power Up Reset
Brownout ($V_{CC} < 4.5V$) Reset
Power Fall Warning
Battery Switchover for CMOS RAM
CMOS RAM and EEPROM Write Protect
Watchdog Timer (Software Monitor)

While the complete microprocessor support circuit in a consumer product might be just a capacitor connected to the Reset pin of the microprocessor, industrial systems often have requirements for precise monitoring of the power supply to generate a high quality Reset signal, and often have some type of non-volatile storage of data while the 110/220VAC power is not available.

An additional circuit that is now common only in high reliability requirements is a watchdog timer. A watchdog timer monitors the operation of the microprocessor by looking for transitions on an I/O pin. If the microprocessor malfunctions or a software bug causes the microprocessor to get stuck in an endless loop, the watchdog timer issues a Reset pulse to the microprocessor, forcing it to restart from the power-up condition.

MAX690/691 MICROPROCESSOR SUPERVISORY CIRCUIT

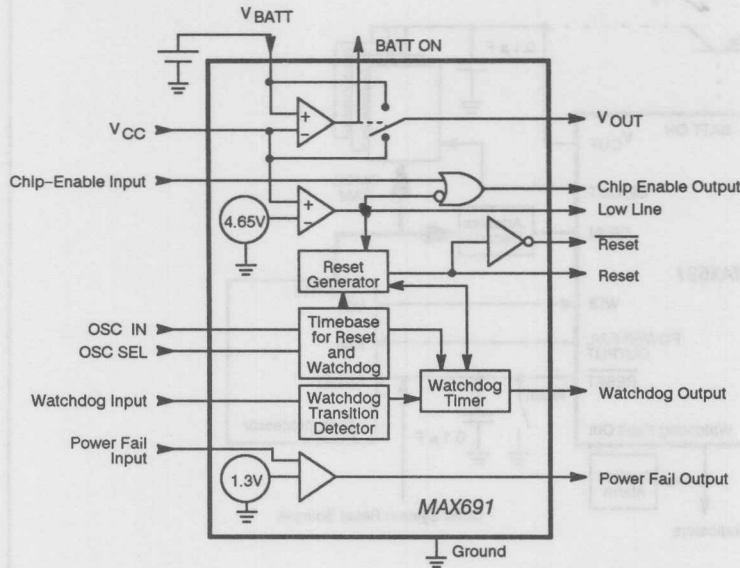
Power-On Reset, Voltage Monitor, Watchdog Timer, Battery Switchover, and Power Fail Detect
No External Components Required
1uA Maximum Standby Current in Battery Backup Mode
Precision Reset Voltage Thresholds
 --4.75 Max, 4.5V Min, 100mV Hysteresis
Reset Generated for Powerup, Brownouts, and Watchdog Timeout
Onboard Delay Timer for Reset Pulse
Watchdog Timeout Periods
 100ms and 1.6 Sec Interval
 10ms to 100+ seconds with External Clock

OUTPUT VOLTAGE vs. OUTPUT CURRENT
FROM V+ TO V-



Output Current (mA)
0 1 2 3 4 5 6 7 8 9 10

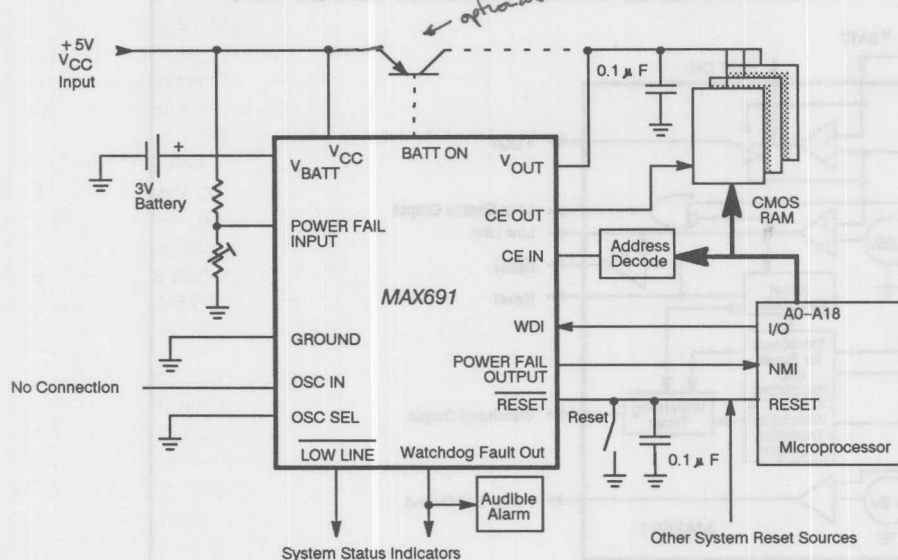
MAX691 BLOCK DIAGRAM



The MAX691 contains several distinct functions that are related only by the fact that they are popular supervisory functions for microprocessor circuit boards. Maxim's goal during the definition of the MAX691 was to combine into one IC as many of these support func-

tions as feasible. To that end, the MAX691 has a battery switchover circuit, a precision 4.65V threshold detector, a 1.3V threshold detector, and timers for both a reset generator and a watchdog circuit.

TYPICAL MAX691 APPLICATION



This circuit uses the many features of the MAX691 to reduce the number of components needed for microprocessor support circuitry.

The Batt On output drives an optional external PNP transistor, increasing the available current on the battery-backed-up power bus to 200mA. Without any help from an external PNP transistor, the MAX691 internal battery switchover circuit will supply 50mA of current with 100mV voltage drop.

The Chip Enable gating circuit inhibits write cycles to the CMOS RAM or realtime clock while VCC is below 4.65V, thereby preventing the microprocessor from corrupting the data in RAM during power-up, power-down, and brownouts.

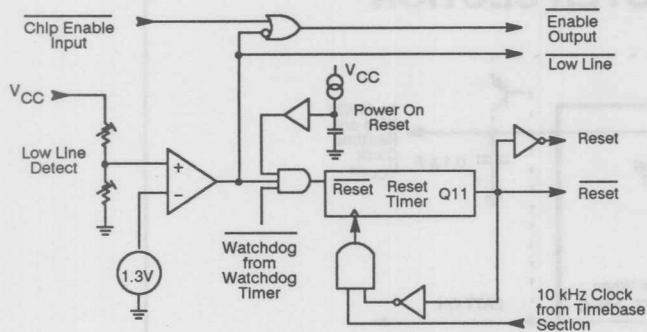
This particular configuration, with Osc Sel pin grounded and Osc In pin floating, ensures a minimum Reset pulse width of 25ms and also sets the watchdog timeout period to 50 milliseconds during normal operation and 200 milliseconds immediately after any reset. If the microprocessor fails to strobe the watchdog input pin (labeled WDI

in the diagram) within this timeout period, the MAX691 will issue a Reset pulse to the microprocessor. The Watchdog Fault Output drives an audible alarm to alert the user that a problem has occurred.

The MAX691 reset pin drives the open collector wired OR reset bus, while also supplying the pullup current needed by the open collector outputs connected to the bus. A manual reset switch and 0.1μF capacitor on the reset bus allow a manual reset of the system.

This circuit shows an alternate method of generating a power fail warning signal, using only the +5V input. The two external resistors connected to the Power Fail Input set up a power fail warning threshold of 4.8V. The processor has the short period as VCC falls from 4.8V down to 4.65V to save any data into RAM. When VCC falls to 4.65V the MAX691 pulls the Reset output low, disabling the microprocessor. The MAX691 CE Output goes high when VCC is below 4.65V, preventing any corruption of battery backed up CMOS RAM data during power-up, power-down, and brownout conditions.

MAX691 RESET BLOCK DIAGRAM

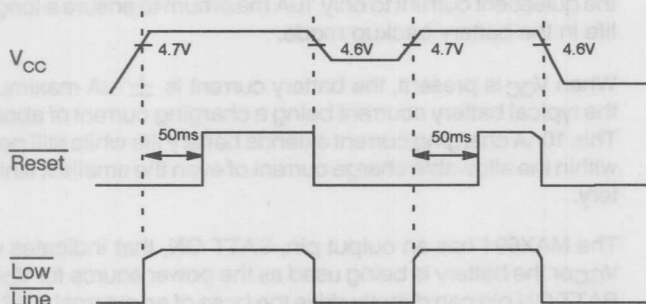


Although shown on the overall MAX691 block diagram as a comparator with a nominal 4.65V reference, the actual internal circuit uses a 1.3V bandgap reference, dividing the 5V nominal V_{CC} to 1.3V using internally trimmed resistors. The Reset outputs are guaranteed to be active whenever the V_{CC} voltage is less than 4.5V. A 50ms timer built from a 10kHz oscillator and an 11 stage divider keeps the Reset outputs active until V_{CC} has remained above 4.75V continuously for 50ms. (See Reset Timing Diagram below.) The 4.65V nominal detection level is precisely set during wafer testing by using fusible metal links to adjust the ratio of the internal resistive divider at the input of the voltage comparator. The 4.65V nominal voltage detector has a guaranteed 4.5V minimum threshold, and a 4.75V maximum guaranteed threshold, with hysteresis of 100mV typical.

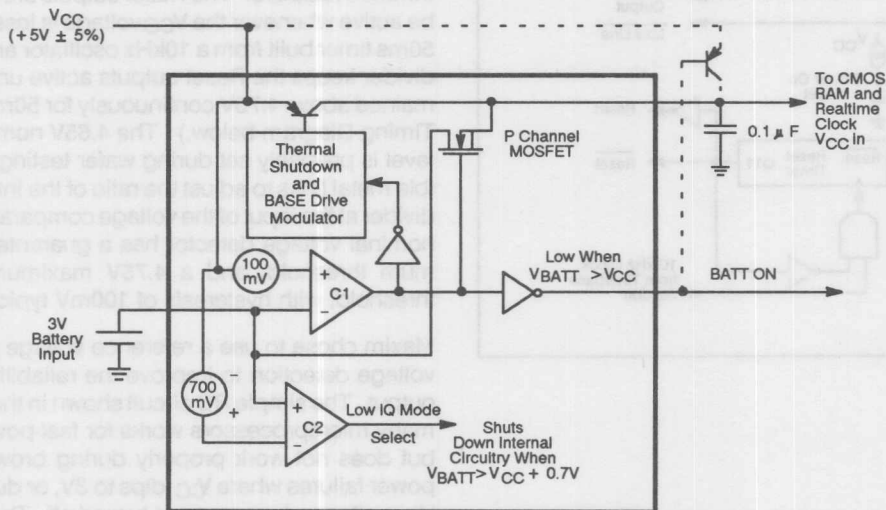
Maxim chose to use a reference voltage and a precise voltage detection to improve the reliability of the reset output. The simple RC circuit shown in the data sheet of many microprocessors works for fast power up cycles, but does not work properly during brownouts, partial power failures where V_{CC} dips to 3V, or during the fall of V_{CC} voltage when power is turned off. The MAX691 Reset output goes low whenever the +5V V_{CC} input is below 4.5V minimum. The MAX690/691 Reset output remains active for 50 milliseconds after the V_{CC} has gone above 4.75V, ensuring that the microprocessor receives the minimum reset pulse width needed for a reliable reset.

In addition to triggering the Reset output, the MAX691 voltage detector also forces the Chip Enable output to a high state whenever the +5V input is at an invalid level. The Chip Enable output can directly drive the Write, $\overline{CS}$, or $\overline{CE}$ pins of CMOS RAMs in systems that can accommodate the 100ns maximum propagation delay of the MAX691 Chip Enable output. For higher speed use the Low Line output to control an external high speed gate.

MAX691 RESET TIMING



MAX691 BATTERY SWITCHOVER SECTION



The voltage comparator C1 compares the V_{CC} input with the battery voltage. As long as V_{CC} is higher than the battery voltage, the MAX690/691 connects V_{CC} to the battery-backed-up output terminal, V_{OUT} , through a saturated PNP transistor. The base drive modulation circuitry keeps the PNP in saturation while minimizing the base current. When V_{CC} falls to within 100mV of the battery voltage, the base drive to the PNP is turned off, and a 750 ohm MOS switch connects V_{CC} to the battery. The MOS transistor minimizes the voltage drop at the sub-milliampere currents normally used by battery backed-up circuitry.

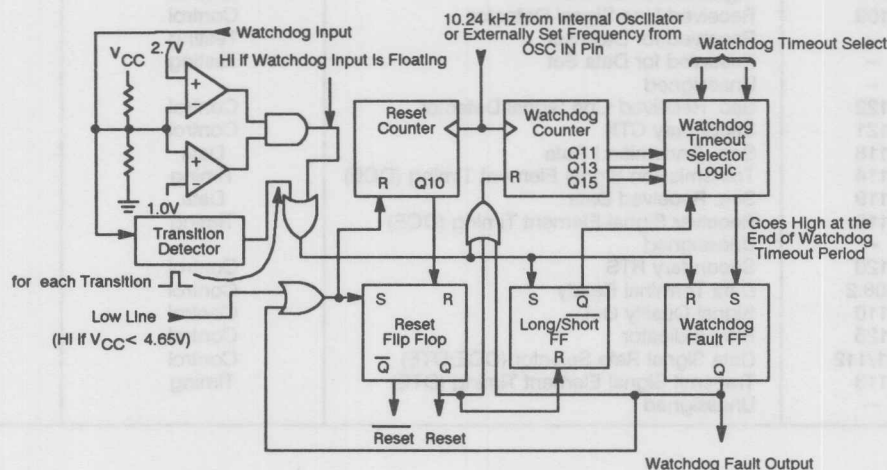
When V_{CC} falls about 700mV lower than V_{BATT} , the comparator C2 activates the MAX690/691 low quiescent current state, shutting off circuitry not needed in the battery backup mode, thereby lowering

the quiescent current to only 1uA maximum to ensure a long battery life in the battery backup mode.

When V_{CC} is present, the battery current is $\pm 1\mu A$ maximum, with the typical battery current being a charging current of about 10nA. This 10nA charging current extends battery life while still being well within the allowable charge current of even the smallest lithium battery.

The MAX691 has an output pin, BATT ON, that indicates whether V_{CC} or the battery is being used as the power source for V_{OUT} . The BATT ON pin can directly drive the base of an external PNP transistor if greater than 50mA of output current is needed to operate the battery-backed-up power bus during normal operation.

WATCHDOG TIMER LOGIC (Simplified Block Diagram)



If the Watchdog Input (WDI) is left floating, two internal resistors bias the input at an invalid logic level and the two internal voltage comparators to disable the watchdog timer.

In a system that does use the watchdog timer, the WDI pin is driven by an I/O line that is periodically strobed. Each time the microprocessor strobes the WDI pin, the MAX691 resets the watchdog timer. If, however, the microprocessor system ceases to strobe the WDI pin, the watchdog timer reaches its maximum programmed time period and a "watchdog fault" occurs. The MAX691 then issues a 50 millisecond long reset pulse, and the MAX691 Watchdog Fault

output goes high. The MAX691 will periodically pulse the Reset line low, and the MAX691 Watchdog Fault output will remain high until the Watchdog Input (WDI) is strobed by the microprocessor.

A variety of timing options allow the setting of the watchdog timeout period to either 100ms or 1.6sec using internal timing, the use of an external clock to set the time period, or adjustment of the timeout periods by the addition of an external capacitor to set the frequency of the onboard oscillator. These timing adjustments also affect the minimum width of the Reset output.

Table 1. Circuit Definitions/Pin Assignments for RS-232/V.24

PIN	EIA RS-232 CIRCUIT	CCITT V.24 CIRCUIT	DESCRIPTION	TYPE	SOURCE
1	AA	101	Protective Ground	Ground	Ground
2	BA	103	Transmitted Data	Data	DTE
3	BB	104	Received Data	Data	DCE
4	CA	105	Request To Send	Control	DTE
5	CB	106	Clear To Send	Control	DCE
6	CC	107	Data Set Ready	Control	DCE
7	AB	102	Signal Ground/Return	Ground	Ground
8	CF	109	Received Line Signal Detector	Control	DCE
9	—	—	Reserved for Data Set	Testing	
10	—	—	Reserved for Data Set	Testing	
11	—	—	Unassigned		
12	SCF	122	Sec. Received Line Signal Detector	Control	DCE
13	SCB	121	Secondary CTS	Control	DCE
14	SBA	118	Sec. Transmitted Data	Data	DTE
15	DB	114	Transmission Signal Element Timing (DCE)	Timing	DCE
16	SBB	119	Sec. Received Data	Data	DCE
17	DD	115	Receiver Signal Element Timing (DCE)	Timing	DCE
18	—	—	Unassigned		
19	SCA	120	Secondary RTS	Control	DTE
20	CD	108.2	Data Terminal Ready	Control	DTE
21	CG	110	Signal Quality Det.	Control	DCE
22	CE	125	Ring Indicator	Control	DCE
23	CH/CI	111/112	Data Signal Rate Selector (DCE/DTE)	Control	DTE/DCE
24	DA	113	Transmit Signal Element Timing (DTE)	Timing	DTE
25	—	—	Unassigned		

Table 2. Circuits Commonly Used for RS-232C and V.24 Asynchronous Interfaces

PIN	CIRCUIT	DESCRIPTION
1	Protective Ground	Connect to Earth Ground
2	Transmit Data (TD)	Data from DTE
3	Receive Data (RD)	Data from DCE
4	Request To Send (RTS)	Handshake from DTE
5	Clear To Send (CTS)	Handshake from DCE
6	Data Set Ready (DSR)	Handshake from DCE
7	Signal Ground	Reference Point for Signals
8	Received Line Signal Detector (sometimes called Carrier Detect, DCD)	Handshake from DCE
11	Printer Busy Signal	Handshake from Printer
20	Data Terminal Ready	Handshake from DTE
22	Ring Indicator	Handshake from DCE

Table 3. Summary of RS-232C and V.28 Electrical Specifications

PARAMETER	SPECIFICATION	COMMENTS
Driver Output Voltage		
0 level	+ 5V to + 15V	With 3–7k Ω load
1 level	–5V to –15V	With 3–7k Ω load
Max. output	$\pm$ 25V Max.	No Load
Receiver Input Thresholds (data and clock signals)		
0 level	+ 3V to + 25V	
1 level	–3V to –25V	
Receiver Thresholds RTS, DSR, DTR		
On level	+ 3V to + 25V	
Off level	Open Circuit or –3V to –25V	Detects Power Off Condition at Driver
Receiver Input Resistance	3k Ω to 7k Ω	
Driver Output Resistance		
Power off condition	300 Ω Min.	$V_{OUT} < \pm 2V$
Driver Slew Rate	30V/ μ s Max.	3k $\Omega < R_L < 7k\Omega$; 0pF < $C_L < 2500$ pF
Signalling Rate	Up to 20kbits/sec.	
Cable Length	50'/15 m. Recommended Max. Length	Longer cables permissible, if $C_{LOAD} \leq 2500$ pF